Electronic technology

principal contents

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SEPTEMBER 1961 VOLUME 38 NUMBER 9

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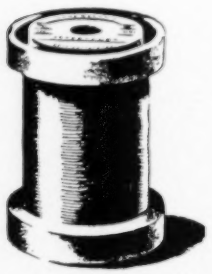

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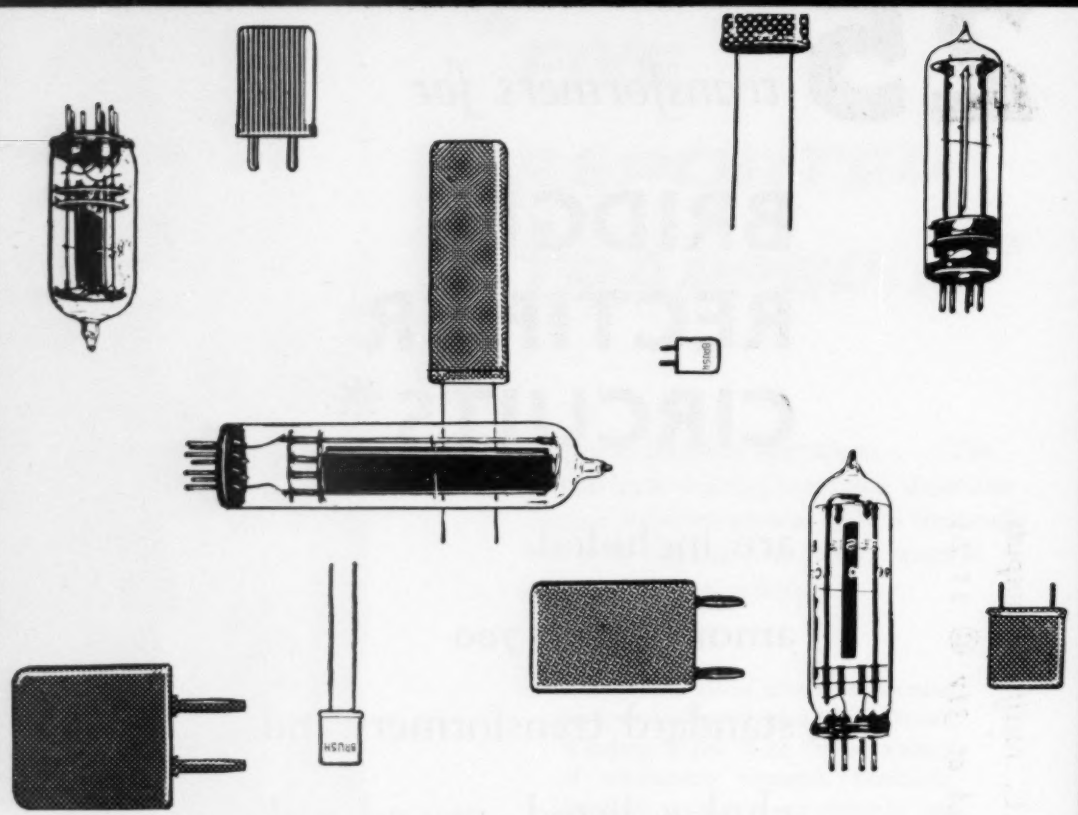

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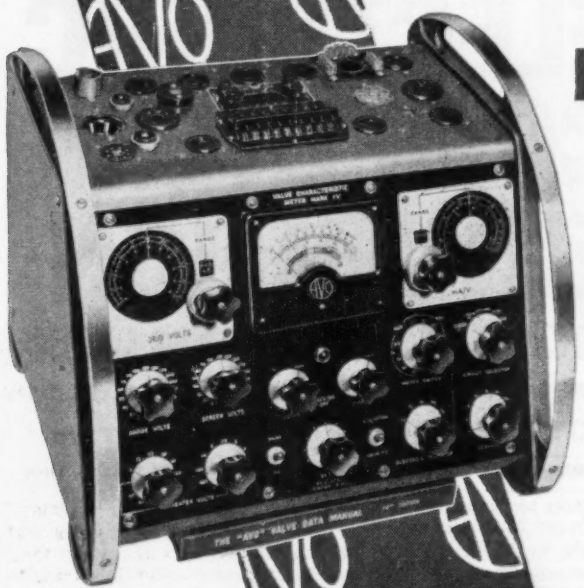

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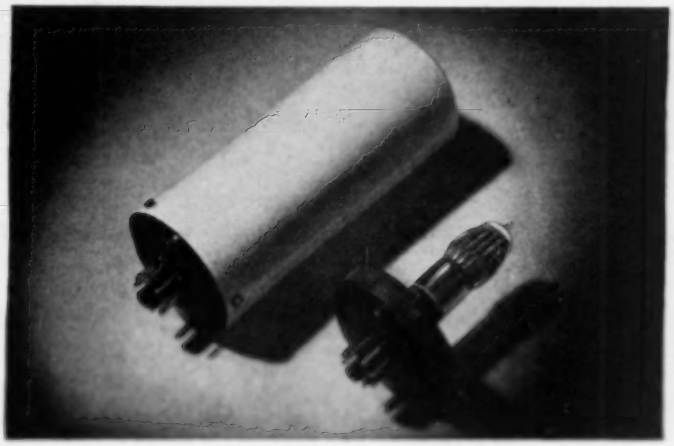

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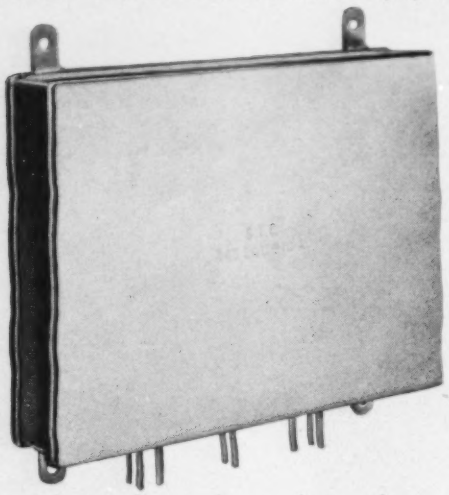

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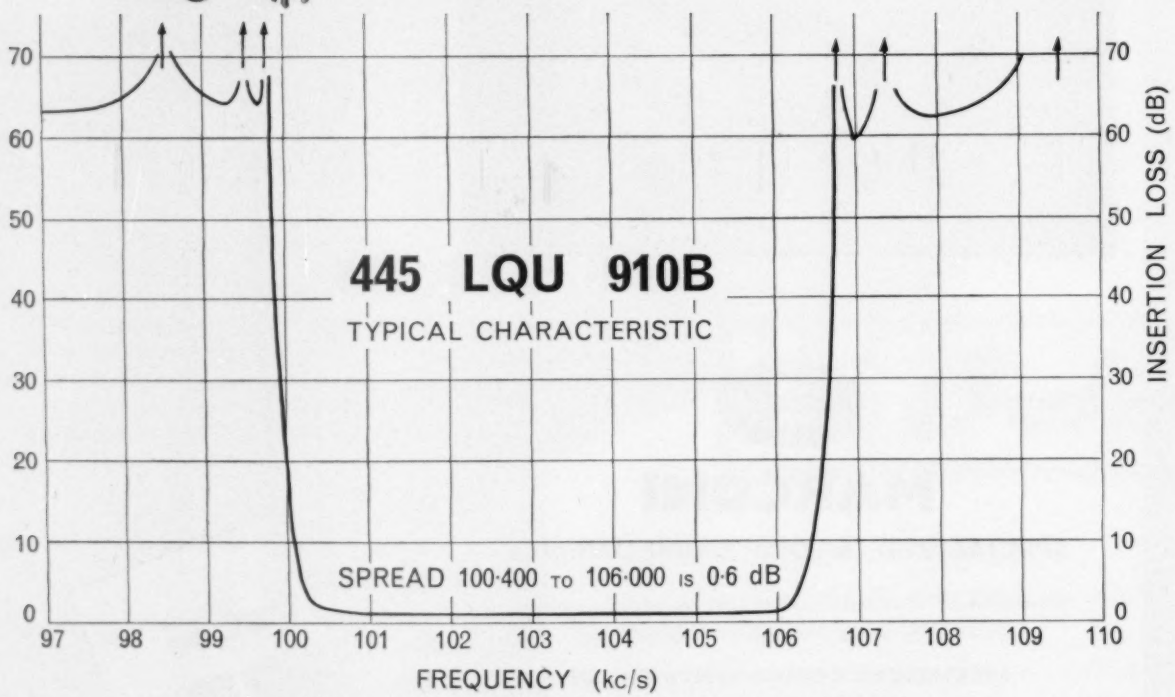

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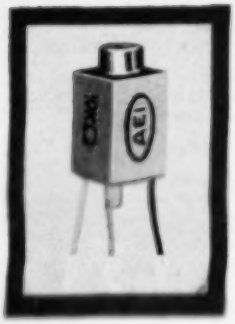

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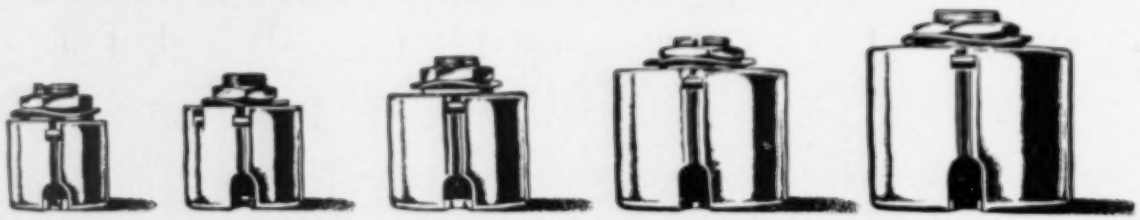

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VOLUME 38 NUMBER 9

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Mathematics

THE use of mathematics in all branches of physics and engineering has been increasing for many years. This is because it is a labour-saving tool which enables a lot of hard thinking to be avoided. Put a problem into mathematical form, and its solution may entail no more than the routine manipulation of equations.

Mathematics may have its own problems, of course, but those are the province of the mathematician. Physicists and engineers tend to become proficient in certain sections only of mathematical knowledge, in those which help them in their real work.

The use of mathematics as a tool is now a normal procedure in all problems involving the relations of quantities, and we all use it as such in greater or lesser degree. We started doing so with the problems of school algebra and over the years we have built on this foundation in our progression to higher mathematics. It is, therefore, very deeply ingrained in all of us that algebraic symbols stand for quantities and represent numbers.

Ordinary algebra is not the only kind of algebra, however. There is an algebra, often called Boolean algebra, of symbolic logic, which has different rules and in which the symbols have no relation at all to quantity. It was originally devised to express logical relations in the form of mathematical equations and so to enable the answers to logical problems to be obtained by the manipulation of equations. It is thus a labour-saving tool in a different sphere of thought.

Its importance to us lies in the fact that it can be applied to the solution of switching problems and to the design of some of the elementary circuit units from which computers are built. In spite of this, it is not generally understood and we therefore include in this issue an article dealing with its elements.

We do not expect that this will make every reader immediately proficient in it! No one will become so unless he is prepared to undertake the hard grind of a lot of practice, but one always has to learn how to use a new tool. Two things we do emphasize. If one wants to understand Boolean algebra the first thing is to put ordinary algebra right out of one's mind, which is difficult, and the second is to remember all the time that the symbols do not stand for quantities, which is very hard indeed.

Transistor Current Gain at U.H.F.

MEASUREMENT TECHNIQUES

By B. N. Harden*, M.Sc., A.M.I.E.E.

SUMMARY. *A coaxial-line system and a twin-channel comparator are used to determine amplitude and phase variations of current gain. Results in the frequency range 300–800 Mc/s are compared with values obtained by a bridge method.*

Various methods[1,2,3] have been adopted for the measurement of the amplitude and phase variations under small-signal conditions of the common-base current gain (α) of transistors at frequencies up to about 300 Mc/s. Other methods[4,5,6] give the amplitude ($|\alpha|$) but not phase, over various frequency ranges up to 1,000 Mc/s and it is possible, within limitations, to deduce the phase from $|\alpha|$ and the amplitude of the common-emitter current gain[7]. The Transfer-Function Bridge[8] enables both amplitude and phase of α to be measured at frequencies up to about 1,500 Mc/s.

The method described here is an extension to u.h.f. of the twin-channel technique used previously[1] at frequencies up to 210 Mc/s. The transistor is mounted in a coaxial-line system and with the equipment in operation at present, amplitude and phase variations of α may be obtained at frequencies from about 300 to 800 Mc/s.

The transfer-function bridge gives values of the short-circuit common-base current gain ($-h_{fb}$)†, whereas in many of the other methods, the 'near-short-circuit' current gain is determined and some allowance for the departure of the termination from short-circuit conditions may be needed to find h_{fb}. The modifications usually involve a knowledge of either the equivalent circuit of the transistor or its input and output impedances as well as the value of the termination, although if the terminating impedance is low, these changes may often be ignored. The twin-channel technique is based on a comparison, in both amplitude and phase, of voltages developed across small impedances in the emitter and collector circuits so that it inherently needs allowance for the terminal impedance if adequate sensitivity is to be obtained; this is particularly relevant in the u.h.f. region.

When a component such as a transistor with wire leads close together is inserted in a coaxial-line system, its measured properties may be appreciably affected by stray impedances at the discontinuities in the coaxial lines. One aim has been to compare experimental results obtained by two different methods (twin-channel and transfer-function bridge) using transistor mounts of different types. The mount for the twin-channel equipment is of simple construction and complex analysis of the effects of strays has not been introduced.

The twin-channel equipment is described and some results obtained with it are compared with others observed on the bridge after allowance for the difference in the quantities actually measured, as discussed above. Although the present equipment does not give h_{fb} directly in the u.h.f. region for many types of transistor, it does provide information of the effect on current gain of departure from short-circuit conditions and gives indirect confirmation of current gain values obtained by other methods.

Method of Measurement

Two circuit arrangements are suitable for the measurement of α as shown in Fig. 1, using either emitter or base injection. If the transistor is represented by its two-port impedance parameters (z_i, z_r, z_f, z_0), z_L is the terminating impedance and z_G the generator impedance, i_e and i_c the emitter and collector currents respectively, then for emitter-injection [Fig. 1 (a)],

$$\alpha = \frac{i_c}{i_e} = \frac{z_f}{z_0 + z_L} = \frac{-h_{fb}}{1 + \frac{z_L}{z_0}} \quad .. \quad .. \quad .. \quad (1)$$

and for base injection [Fig. 1 (b)],

$$\alpha = \frac{i_c}{i_e} = \frac{z_f - (z_L + z_i)}{z_0 + (z_L - z_r)} = \frac{-h_{fb} - \left(\frac{z_L + z_i}{z_0}\right)}{1 + \left(\frac{z_L - z_r}{z_0}\right)} \quad .. \quad .. \quad .. \quad (2)$$

The determination of h_{fb} from the measured ratio of i_c/i_e involves in Equ. (1) a knowledge of z_L/z_0 but in Equ. (2), z_i/z_0 and z_r/z_0 are also required. In many cases, z_L/z_0 may be made negligible by suitable choice of z_L and the other ratios may be ignored, but in the u.h.f. region this may not be possible.

In earlier measurements[1] at frequencies up to 210 Mc/s, emitter injection was used and the terminating impedances z_L were small resistors (50 Ω); the complex current ratio i_c/i_e in this case was equal to the complex ratio of the voltages existing across the resistors. The

* *D.S.I.R. Radio Research Station, Slough.*

† In this paper the current gain with an arbitrary collector-base termination is designated by α and the short-circuit current gain by $-h_{fb}$.

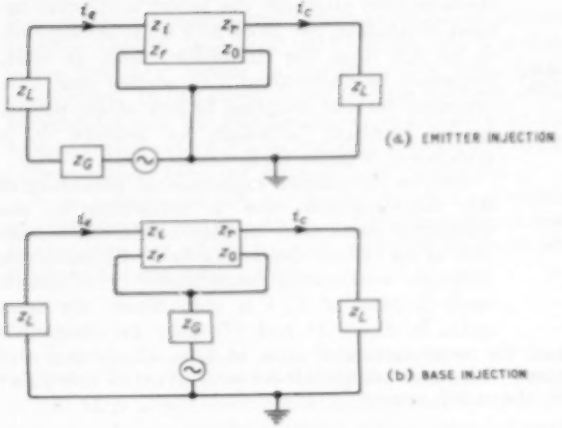

Fig. 1. *Circuits for the measurement of α with the transistor represented by its two-port impedance parameters*

voltages were applied to two receivers with identical amplitude and phase responses and the outputs displayed on a c.r.o., the amplitude and phase relationships between them being determined from the shape of the elliptical trace.

In the present equipment, the terminating impedances z_L are formed from two lengths of lossy coaxial-cable which provide loads of 72 Ω impedance in both emitter and collector circuits over a wide range of frequencies. These are connected to the coaxial transistor mount by short lengths of air-filled coaxial-line of the same characteristic impedance. The current ratio may then be found from the ratio of the voltages existing at specific positions in the lines, chosen in accordance with the configuration in use. The voltages are sampled by probes and the signals passed to two identical receivers.

The following relationships apply in the coaxial-line system:

Emitter Injection

If z_{in} is the input impedance of the transistor with collector load z_L, the complex voltage at a point P in the coaxial-line a distance d towards the emitter termination from the transistor input is given by

$$v_P = \left(z_{in} \cos \frac{2\pi d}{\lambda} + j\, z_L \sin \frac{2\pi d}{\lambda} \right) i_e$$

where λ is the wavelength.

When $d = \lambda/4$, $v_P = j\, i_e\, z_L$.

The voltage at a point Q a distance s towards the collector load from the transistor output is given by

$$v_Q = i_c\, z_L \exp -j \frac{2\pi s}{\lambda}$$

The ratio of the voltages is then

$$\frac{v_Q}{v_P} = \frac{i_c}{i_e} \cdot \frac{\exp\left(-j \frac{2\pi s}{\lambda}\right)}{j}$$

If the transistor is replaced by a direct connection between the inner conductors of the coaxial-line sections, the ratio of the voltages at P and Q is then given by

$$\frac{v_{QS}}{v_{PS}} = \frac{\exp\left(-j \frac{2\pi s}{\lambda}\right)}{j}$$

so that $\alpha = \frac{i_c}{i_e} = \frac{v_Q}{v_P} \cdot \frac{v_{PS}}{v_{QS}}$

Under the same conditions, when probes with coupling factors K_P and K_Q are placed at P and Q and the signals from them passed through receivers with

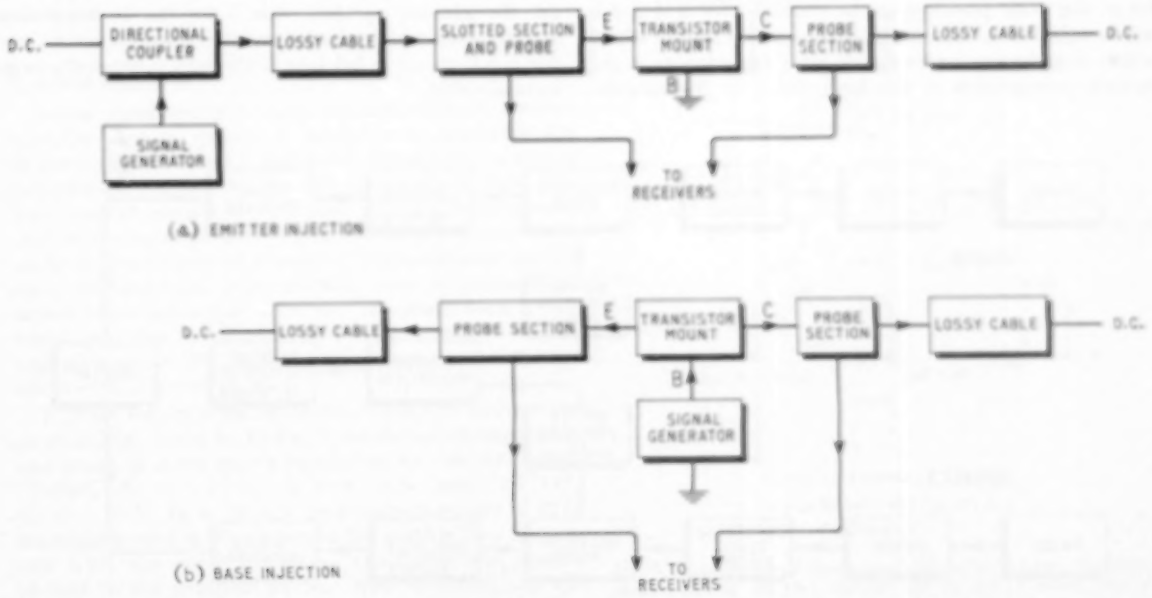

Fig. 2. *Schematic diagrams of coaxial systems.* (*Transistor connections in mount: E emitter, B base, C collector*)

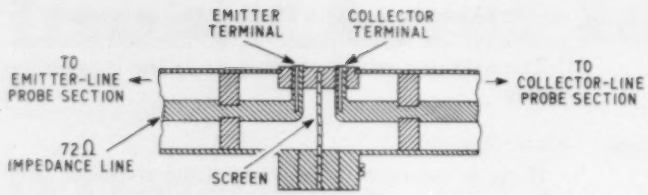

Fig. 3. *Coaxial transistor mount. A schematic cross-section through the emitter and collector terminals; the base terminal is in the plane of the screen*

gain factors G_P and G_Q adjusted so that the ratio of the outputs is unity (as indicated by a straight line at 45° on the c.r.o. display) then

$$v_{QS}\, K_Q\, G_Q / v_{PS}\, K_P\, G_P = 1$$

With the same probe and receiver settings and the transistor in circuit, the ratio displayed is $v_Q\, K_Q\, G_Q / v_P\, K_P\, G_P$ which equals $v_Q\, v_{PS} / v_P\, v_{QS} = i_c / i_e = \alpha$

Base Injection

With the same notation as used above

$$v_P = -i_e\, z_L \exp \frac{-j2\pi d}{\lambda}$$

$$\text{and } v_Q = i_e\, z_L \exp \frac{-j2\pi s}{\lambda}$$

$$\text{so that } \frac{v_Q}{v_P} = -\frac{i_c}{i_e} \cdot \exp \frac{-j2\pi (s-d)}{\lambda}$$

If a balancing procedure similar to that for emitter-injection is followed, the ratio of the responses from probes at P and Q may be directly related to the ratio i_c/i_e, that is, to α.

Restrictions imposed on the experimental procedure by the required positioning of the probes are simpler for the base-injection case. At all frequencies the probes may be at the same positions in the coaxial-lines and if d is made equal to s, a balanced arrangement results which assists in reducing errors due to stray impedances in the mount, mismatches in the load, etc. In the emitter-injection case although the probe at Q may be fixed in position, the probe at P must be adjusted to be $\lambda/4$ from the transistor input at each frequency of observation. In neither case is it necessary for the coupling factors of the probes to be identical, although in practice it is convenient to make them nearly equal.

Despite the simpler experimental procedure of the base-injection case in measuring α, the additional restrictions imposed on the determination of h_{fb} reduce this advantage. By using two different terminating impedances z_L of known ratio (a ratio of 2 : 1 is convenient) the term z_L/z_0 in Equs (1) and (2) may be eliminated but the terms z_i/z_0 and z_r/z_0 in Equ. (2) remain and these may not be negligible for some types of transistors in the u.h.f. region.

Equipment

Coaxial-Line System

Schematic diagrams of the emitter- and base-injection coaxial systems are shown in Fig. 2. Quite short lengths (about 4 metres) of lossy coaxial cable, giving an attenuation of 15 to 20 dB at the lowest frequency, are sufficient to act as satisfactory loads to the test transistor and the d.c. supplies to emitter and collector are applied through the centre conductors of these cables, eliminating any by-pass capacitors and blocking resistors. The use of a directional-coupler permits easy separation of the u.h.f. and d.c. sources.

The test transistor is mounted perpendicularly to the axis of the coaxial-line sytem through a cut-away section of the outer conductor as shown schematically in Fig. 3. The emitter and collector leads are inserted into short split tubes on the ends of the inner conductors and a screen, connected to the outer conductor, is situated symmetrically between them. Another short tube for the base lead is appropriately positioned on this screen for the emitter-injection case. In the base-injection system, the base terminal is isolated from the screen and the u.h.f. signal is fed into it through a small-diameter coaxial-cable.

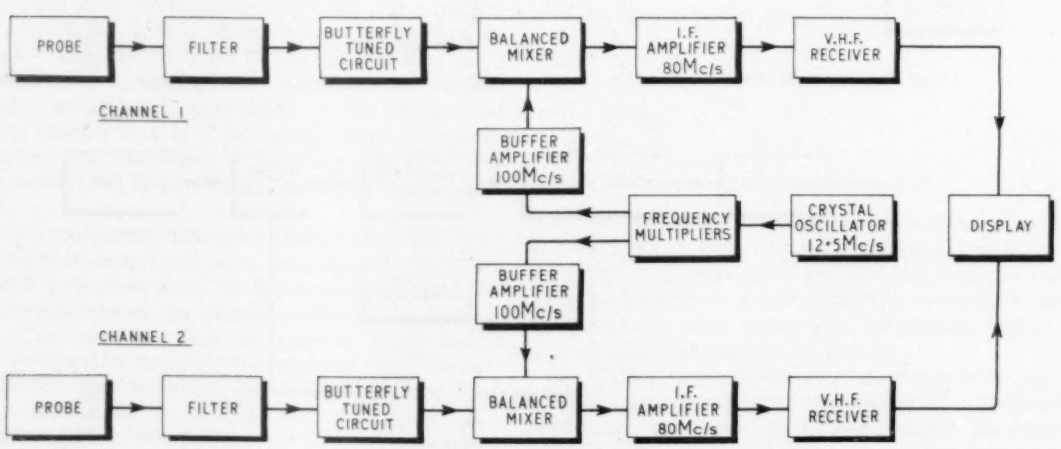

Fig. 4. *A schematic diagram of the receivers*

The dimensions of the mount are adjusted to suit the particular type of transistor under investigation. Several types of mounts have been used but the one described above which is simple to construct has been found satisfactory. In this mount a compromise has been made between keeping the lead lengths short and keeping the transistor case, which is often connected internally to the collector, away from the earthed outer conductor.

Probe penetration into the coaxial system is adjustable and coupling factors of −15 to −20 dB have been found to give adequate sensitivity with negligible effect on the observed voltage distributions.

Receivers

Twin converter-units were constructed to be used with the v.h.f. receivers covering the frequency range to 210 Mc/s which were available from earlier measurements[1] and were already provided with separate controls for amplitude and phase adjustment of each channel. A schematic diagram of the circuit is shown in Fig. 4.

For stability a crystal-controlled local-oscillator signal is desirable. The output from a 12·5 Mc/s crystal oscillator is frequency multiplied to 100 Mc/s. Buffer amplifiers eliminate interaction between the two channels and sufficient signal at 100 Mc/s is injected into balanced mixers to permit the harmonics to be used for frequency changing the u.h.f. signals to an i.f. of 80 Mc/s. Thus input signals at a number of frequencies, 280, 320, 380, 420, etc, Mc/s may be received.

The u.h.f. input signals from the probes are loop coupled to wide-range tuned circuits of the butterfly type connected to the balanced mixers. High-pass filters prevent any stray radiation from the 100 Mc/s units from entering the coaxial system; the balanced mixers and selectivity of the tuned circuits give sufficient rejection of harmonics of the 100 Mc/s local-oscillator signal.

Amplitude-and phase-balancing controls are provided in addition to those already available in the v.h.f. receivers.

Typical Results

Initial measurements were commenced with a base-injection system before a bridge was available for determining in the u.h.f. region the values of z_t, z_0 and z_r required to evaluate h_{fb} for this system from Equ. (2). Two coaxial systems of 72 Ω and 36 Ω impedance were used to eliminate the ratio z_L/z_0 and measurements were made on two types of transistor, a germanium micro-alloy diffused-base type 2N502, and a germanium diffused-base mesa type 2N1142. However, since it was found that the ratios z_t/z_0 and z_r/z_0 were not always negligible above 300 Mc/s no further measurements were taken with this system.

Typical results with an emitter-injection system are given in Figs. 5 and 6. In Fig. 5 the measured amplitude and phase of α are shown separately for two transistors (2N502, $V_C = -10$ V, $I_E = 3$ mA and 2N1142, $V_C = -10$ V, $I_E = 10$ mA) in a coaxial system of 72 Ω impedance, and in Fig. 6 results for another two transistors with the same operating conditions are shown plotted in the complex plane. Also indicated are the values of $-h_{fb}$ obtained with a transfer-function bridge

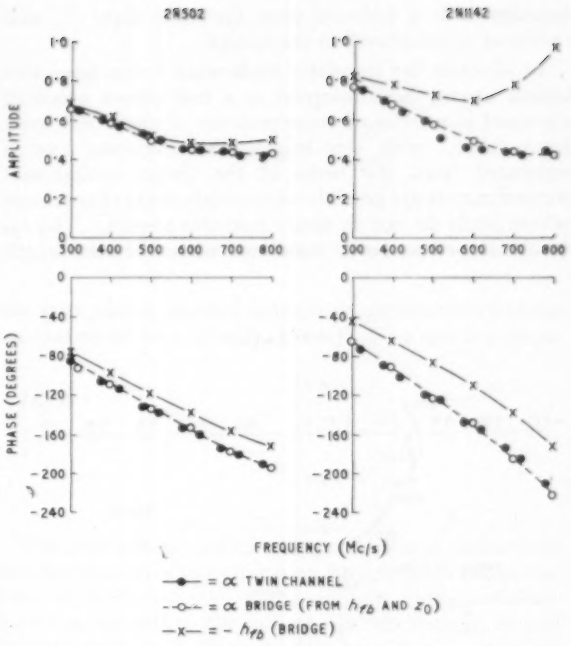

Fig. 5. *A comparison of the measured amplitude and phase values*

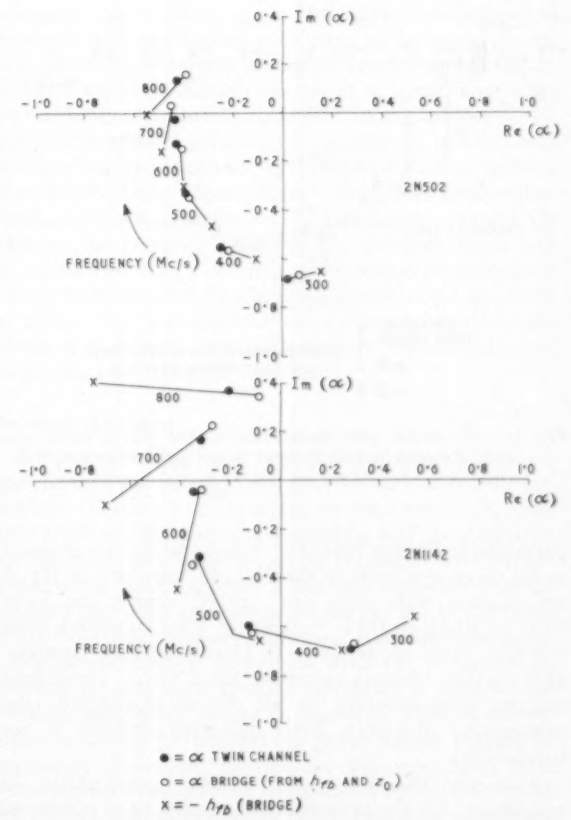

Fig. 6. *A comparison of the measured values of α. Corresponding points are shown linked at 100-Mc/s intervals and the loci of current gain versus frequency are given by curves through appropriate points*

together with α deduced from h_{fb} using Equ. (1) and values of z_0 measured on the bridge.

In all cases the transistor leads were 5 mm long, this length having been adopted as a convenient practical standard in a series of measurements of which this work forms part; with this length, the transistor case is separated from the body of the bridge mount and measurements are possible with certain types of transistor whose leads do not fit easily into this mount. The h_{fb} results show resonance associated with this lead length

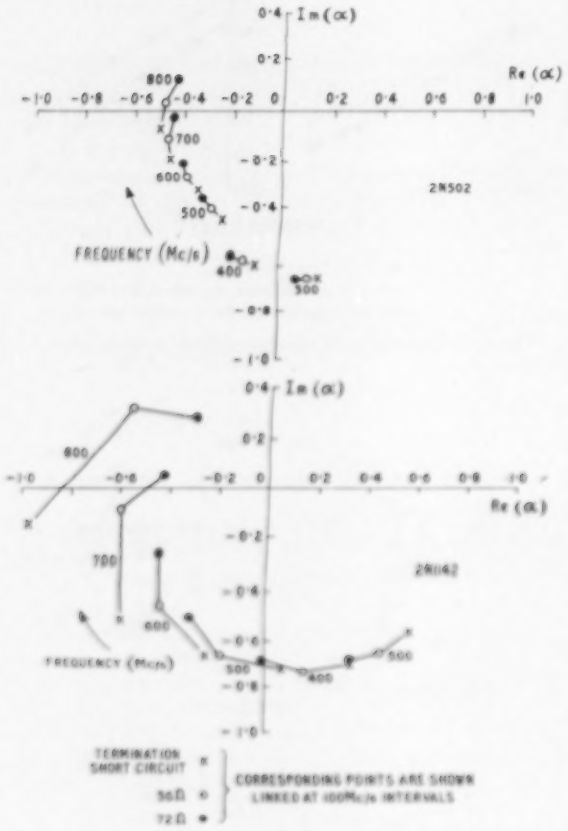

Fig. 7. *The current gain characteristics. (The loci of current gain versus frequency are given by curves through appropriate points)*

particularly for the 2N1142. For instance, the changes in h_{fb} when the leads of the 2N1142 transistor of Fig. 5 are pushed right into the bridge mount are from 0·07 $-$ *j*0·73 to 0·11 $-$ *j*0·68 at 500 Mc/s and from $-$0·97 $-$ *j*0·15 to $-$0·59 $-$ *j*0·23 at 800 Mc/s. Appreciably smaller changes occur for the 2N502. Since lead lengths were constant in all the measurements, the comparison of results from different methods is not invalidated.

Although there is some scatter, particularly in amplitude, for the twin-channel results there is general agreement between measurements by the two methods. Since the transistor mounts are of entirely different construction and the wire ends of the transistor introduce discontinuities in the coaxial system, the agreement shows that the neglect of the resultant strays is not serious in the frequency range covered. Differences in detail are however most likely to be due to these strays and accurate evaluation would involve a complex analysis of the coaxial system.

The emitter-line probe position at $\lambda/4$ in front of the transistor input may ideally be found by observing the voltage distribution along the slotted section of line when it is either short- or open-circuited. Since in practice it is convenient to obtain approximations to these conditions by connecting the emitter and base terminals of the mount with a short length of wire or by leaving them open, some variations in results obtained in the two cases may be expected. With the 72-Ω impedance system differences in phase were small but for amplitude, differences varying with frequency were observed; these reached a maximum of about 7% between 600 and 700 Mc/s with a 2N1142 transistor but were substantially less over the rest of the frequency range and with 2N502 transistors.

Since with these transistors and operating conditions, large variations in current gain were observed with a short-circuit termination and one of 72 Ω, it is of interest to examine the changes to be expected on halving the terminal impedance to 36 Ω. Fig. 7 shows results for a 2N1142 and a 2N502 calculated from values of h_{fb} and z_0 measured on the bridge and it is evident that even with a terminal impedance of 36 Ω (a value which is often considered to be sufficiently low to give 'near-short-circuit' conditions) appreciable differences in α are obtained in the u.h.f. region.

Conclusions

The twin-channel method of measuring the complex current gain has been extended to the u.h.f. region using coaxial-line techniques. With a transistor mount of simple construction, agreement has been obtained with values determined by a bridge method. The effect on current gain of departure from short-circuit terminal conditions has been shown to be appreciable for some types of transistor at u.h.f.

Acknowledgements

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Symbolic Logic

FOR COMPUTER CIRCUIT APPLICATION

By B. R. Wilkins, B.Sc. (Eng.)

SUMMARY. *The elements of Boolean Algebra and Venn Diagrams are explained, and their applications considered first to verbal statements and then to the electrical description of some switching and computer circuits. Finally, the methods are used in a simple design problem of a type which could be met in a digital computer.*

The starting point in the design of any kind of machinery is the specification setting out the operation which is to be performed by the machine. The designer has available certain components out of which the machine is to be built. In practice it is found, especially in the case of electronic computers, that the specification can best be met by working with a number of sub-units, each of which fulfils a definite and limited function, and using these units as 'building bricks'. The logical design of a computer consists of deciding how these 'bricks' may best be fitted together to satisfy the specification. It is found that the operation performed by each 'brick' (known as a logical operation) can be represented mathematically, and indeed it is this aspect of the problem which gave rise to the name 'logical', because it turns out that the mathematical techniques involved are identical with those devised by the Irish mathematician, George Boole, in 1854 for the symbolic representation of logic.

This mathematical representation is important for three reasons. The first is that it can act as a convenient shorthand notation for expressing the function of any particular section of the design. The second is that circuits can be designed mathematically working from a mathematical representation of the specification. The third is that great simplification is often possible in a complicated design. This simplification is by no means obvious from consideration of the circuit configuration, but can be easily derived by the use of Boolean algebra.

It will be shown later that the complication of the Boolean expressions for a logical circuit is directly related to the complication of the required circuit, and therefore simplification reduces cost, and even if the saving amounts to no more than one or two circuit components in any particular case, it should be remembered that in a digital computer some basic circuits are often repeated hundreds or even thousands of times, so that any saving in the complexity and cost of a unit is worthwhile.

Symbolic Logic

Since Boolean Algebra was originally intended for use in Symbolic Logic, a few words about the historical background of this subject may be of interest.

The problem of reducing arguments and deductions to a mathematical form has been thought about since the time of Aristotle, and many eminent mathematicians have contributed to the subject. It was thought at one time that this would form the basis for a universal algebra whereby all knowledge, including moral and theological truths, could be derived within a single deductive system. Less ambitious, but more successful, attempts have been made to apply mathematical methods to reasoning by such men as Leibnitz, Euler, and, in his more serious moments, Lewis Carroll. More recently Bertrand Russell examined at great length the logical basis of mathematics itself. In the middle of the last century, Venn produced a geometrical method and Boole an algebraic one for the analysis of statements and their implications. Immediately afterwards many other mathematicians contributed, one of whom, named De Morgan, propounded an important theorem which will be mentioned later. All this work remained unnoticed except by specialized logicians for nearly a century, until by about 1950 the applications of these methods to the design of switching and relay circuits and digital computers became generally apparent.

Boolean Algebra

Before embarking on a study of Boolean Algebra we must first rid our minds of all ideas derived from ordinary algebra. The signs and symbols we shall use are in most cases borrowed from ordinary algebra, but the meanings are not. It is only too easy to be led astray by a superficial resemblance between a Boolean equation and an ordinary algebraic one unless we are careful to interpret all symbols in accordance with their Boolean definitions.

Boolean algebra is concerned with the relationships between classes: each symbol represents a class whose members have some feature in common. Thus in talking about a group of people we may choose classes of Spaniards, Germans, Europeans, Painters, Men and Women, and these may be represented by *S*, *G*, *E*, *P*, *M* and *W*. We must notice at once that there is no suggestion of number associated with these symbols: *S* represents not a quantity of Spaniards, but rather the quality of being Spanish.

When examining any set of statements it is important that the universe of discourse should be defined. For example, the statement "No Germans are Painters" (which may be written for brevity "No G is P") may be true for the particular group of people we are talking about, but can obviously not be extended to any other context. This leads us to the idea of the universal class, that is the class which contains every member of the universe of discourse. This class is denoted by "1", while a class which is devoid of members is denoted by "0".

Two classes are said to be equal if they have all members in common: that is A and B are equal if and only if all members of A are also members of B, *and* all members of B are also members of A. The equivalent abbreviated way of saying this is that if All A is B *and* All B is A, then A and B are equal, and we write

$$A = B.$$

I should perhaps remark that the notation in this subject is not standardized, and variations on what I use here will be encountered in many textbooks.

From any given class A we can form the complementary class, which is the class containing all those members of the universe of discourse which are not members of A. This class is called "not A", and is written A'. Thus in our group of people we can form the class of all those who are not Spanish, and this will be denoted by S'. We may now observe that one of the six postulated classes is redundant, because the classes of men and women are complementary, that is $M = W'$ and $M' = W$. Notice however that though the classes S and G are mutually exclusive—that is,

No S is G *and* No G is S

they are not necessarily complementary: there may be members of the group who are neither Spanish nor German.

There are two basic operations which can be performed on classes. The first is the "and" operation: forming the class C consisting of those members of A who are also members of B. This is expressed as

$$C = AB$$

and is spoken of as Boolean multiplication. We must again observe that there is no numerical implication in either the operation or its symbolic representation. If the class of German painters in our group of people is designated class J, then $J = GP$ but clearly the number of members of class J bears no simple relationship to the number of members of either classes G or P.

The second basic operation on classes is the "or" operation. Here we encounter a slight linguistic difficulty. The "and" operation follows quite reasonably from the ordinary way of talking—members of J are members of G *and* members of P—but "or" is used in two different ways, as we may see in two simple examples.

A child with sixpence in his pocket can buy a bar of chocolate *or* a ping pong ball. A man takes a taxi when it is raining *or* when he is in a hurry. The child can pursue either one of his alternatives, but they are mutually exclusive—if he has a bar of chocolate he cannot also have a ping pong ball. The conditions for the man to take a taxi however are not mutually exclusive—it is possible that he may be in a hurry and that at the same time it is raining, and in this case also he will take a taxi. The first type of "or" is called the "exclusive or" (either but not both), and the second is called the "inclusive or" (either or both). The inclusive or is the one taken to define the meaning of "or" in Boolean Algebra. If we form the class D whose members are members of A or of B or of both, then we write

$$D = A + B$$

and this is spoken of as Boolean addition.

The "exclusive or" function can be defined and given a separate symbol, but the operation can be expressed by means of the symbols already defined as

$$E = AB' + A'B$$

Symbols can also be assigned to other operations, but for our purposes these ideas are sufficient, and in computers the basic logical sub-units are normally "and", "or" and "not" circuits.

The choice of multiplication and addition symbols for "and" and "or" operations is deliberate, because it can be shown that Boolean symbols obey the associative, distributive and commutative laws of normal algebra if they are treated as if they were multiplication and addition symbols. It should be noted though that division and subtraction have not been defined, and have no meaning.

From the definitions given above, the following identities emerge:

$$\begin{aligned} AA &= A & A + A &= A \\ A.1 &= A & A + 0 &= A \\ A.0 &= 0 & A + 1 &= 1 \\ AA' &= 0 & A + A' &= 1 \end{aligned}$$

These relationships demonstrate the dual nature of the algebra, and in fact it can be shown to be generally true that in any given identity we may interchange "and" and "or" operations, and interchange "0" and "1", and obtain another identity.

Two formulae which are useful for simplification can be derived from the above equations:

$$\begin{aligned} A + AB &= A \\ A + A'B &= A + B \end{aligned}$$

We may also deduce that if

$$A + B = 0 \quad \text{then} \quad A = B = 0$$

One final formula which is required is the result of applying the "not" operation to a function of more than one variable. We make use of De Morgan's theorem which states that this result is obtained by interchanging "and" and "or" operations, dashing the undashed symbols and undashing the dashed ones.

By means of these relationships, equations obtained from the specification of a design problem can be simplified mathematically before being translated back to a physical circuit[1].

Two examples of the application of Boolean algebra to verbal statements will serve to bring out some useful points.

First consider the statement that all members of class A are also members of class B, or all A is B. Another way of putting this is that class A is entirely contained in class B. We must note that it does not follow that the classes are equal, because although all A is B, we are not

told that all B is A, so that the conditions for equality are not necessarily satisfied. The situation can be expressed in several ways, two of which are

$$AB = A \quad \ldots \quad (1)$$

$$A + B = B \quad \ldots \quad (2)$$

If we applied normal algebraic division and subtraction rules to these equations, we would deduce

$$\text{from (1)} \quad B = 1 \quad \ldots \quad (3)$$

$$\text{and from (2)} \quad A = 0 \quad \ldots \quad (4)$$

This demonstrates that division and subtraction are not valid processes since these deductions are obviously not valid, as may be shown in a specific example. Suppose that the statement is "All Englishmen are Europeans", the universe of discourse being "human beings". Then in the above equations A represents Englishmen, and B represents Europeans, and translating equations (3) and (4) back into words, the deductions are that all human beings are Europeans, and that no Englishmen are human beings.

As a second example, consider the statements

All A is B

All B is C

from which a conclusion is to be drawn.

Now, another way of expressing these statements is

$$AB' = 0 \quad \ldots \quad (5)$$

$$\text{and } BC' = 0 \quad \ldots \quad (6)$$

Applying De Morgan's Theorem to Equ. (5) gives

$$A' + B = 1,$$

and if we multiply this by C', we have

$$A'C' + BC' = C'.$$

But from Equ. (6), $BC' = 0$, and so $A'C' = C'$.

We may apply De Morgan's Theorem again to give

$$A + C = C$$

which we recognize as "All A is C", and this is the required deduction, which is clearly correct.

Venn Diagrams

Another mathematical system of logic which has important applications in computer work is the geometrical system devised by Venn[2]. Venn diagrams can be particularly useful when visualizing a complicated situation.

The basic idea behind Venn diagrams is that classes are represented by areas. Thus, in Fig. 1 (a), the universe of discourse is represented by a square, and the class A is represented by a circle. The area between the circle and the square then must represent A'. Now, if we have three classes, A, B and C, these may be represented by three intersecting circles as shown in Fig. 1 (b), and it will be seen that every combination of A, B and C and their negatives is represented by an area on the diagram. This is shown in Fig. 1 (c). If we wish to show more than three classes, then circles can no longer be used. For four classes, we may use ellipses as shown in Fig. 1 (d). In this the boundary square is not shown. This is normally omitted in practice—the space surrounding the diagram and bounded by the universe of discourse (representing $A'B'C'D'$) is denoted by an extra circle

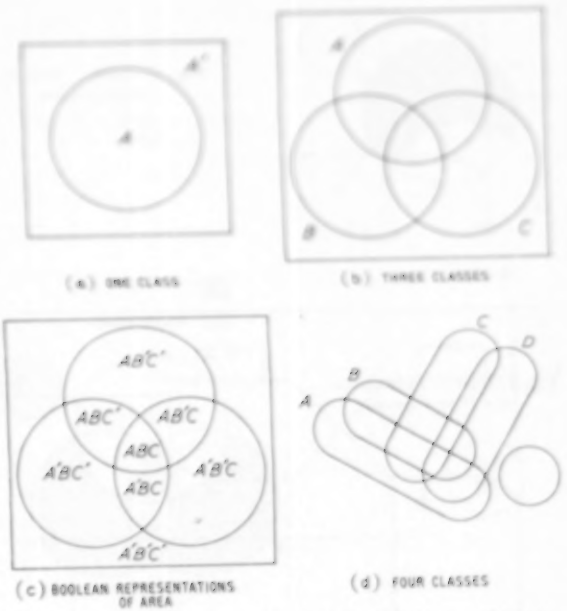


Fig. 1. *Venn diagrams; (a) one class, (b) three classes, (c) Boolean representations of areas, and (d) four classes*

as shown. With more than four classes Venn diagrams can be used, but the curves which have to be used become so complicated that the method is not suitable.

Statements may be represented on the diagram by shading those sub-classes which are empty and marking with an X those sub-classes which are known to have members. One or two examples will serve to make the method clear.

Consider first the example used before:

All A is B

All B is C

These may be put

$$AB' = 0 \text{ and } BC' = 0.$$

If we shade the appropriate areas as in Fig. 2, we see at once that the only part of A which is unshaded is contained in C. Thus we deduce

All A is C.

As a second example, consider the statements

All A is B

Some C is not B

The second statement tells us that the class CB' is not

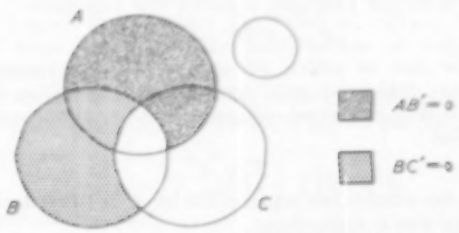


Fig. 2. *Solution of simple problem by Venn diagram*

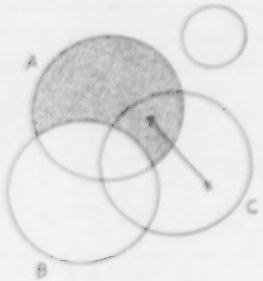


Fig. 3 (*Left*). *Solution of second problem*

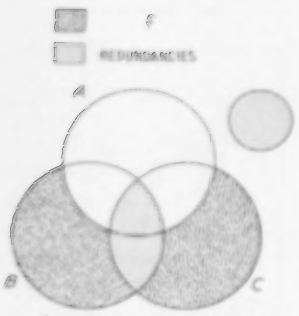


Fig. 4 (*Right*). *Redundancies on a Venn diagram*

empty. This class, however, consists of two sub-classes, and the statement only tells us that at least one of these is occupied. The situation is diagrammed by putting an X in each area, with a line between them to indicate an indecision (Fig. 3). The first statement, as in the previous example, means that the class AB' is empty, so that the appropriate area is shaded. We now find that the ambiguity in the second statement is resolved: since one of the two sub-classes comprising CB' is empty, then the other one must be occupied, and we may deduce from the diagram that

Some C is not A.

Perhaps the most useful application of Venn diagrams is for visualizing ways of simplifying Boolean functions. Groups of sub-classes which can be expressed by a single term are arranged on a Venn diagram in an ordered and easily recognizable manner. This aid to simplification is especially useful when there are redundancies. This is a case which often arises, where some particular sub-classes for some reason may not exist. Such classes are variously referred to as redundancies, irrelevancies, forbidden combinations, or, with a certain pleasing directness, 'don't care' conditions. The point is that if a particular class is known to be empty, then we don't care whether an expression we use includes this class or not.

As a simple example, consider the expression

$$F = A'BC' + A'B'C$$

for the case in which BC and $A'B'C'$ are redundant. We may depict this in a Venn diagram as in Fig. 4, and we see that by taking account of the redundancies we may say

$$F = A'$$

since the whole area represented by A' is either included in F or else is redundant.

This is really equivalent to saying that since the redundant terms are equal to zero, we may add as many of them as we like to an expression without affecting its value. Thus in the expression above, we have put

$$F = A'BC' + A'B'C + A'BC + A'B'C'$$

the last two terms being equal to zero.

A Venn diagram provides a much easier way of deciding how the redundancies may usefully be incorporated than the purely algebraic approach.

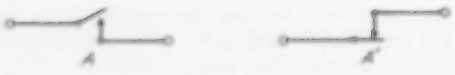


Fig. 5. *Symbols for relays*

Electrical Applications

Before mentioning the application of Boolean expressions to switches and relays, a few words must be said about two-valued logic. This is simply the particular case in which each class is in one of two states, either full or empty. Two classes are then said to be equal if they are in the same state. If the two states are represented by "0" and "1", all the previously stated relationships hold, despite the different interpretation of "1" and of equality. The use of "0" and "1" for these two states is especially convenient in digital computers, where binary digits clearly satisfy the conditions of two-valued logic.

An obvious interpretation of a function with two states is a switch which is either open or closed. Further, we may think of any expression as a network of switches with one terminal at each end. The whole network is then itself a switch, and depending on the states of the intermediate switches there either is or is not a connection between the terminals.

A slight extension of the idea permits the interpretation of dashed functions. If we think of each switch as being operated by a relay, then we may consider simple functions as being normally open switches and dashed functions as being normally closed switches, so that the simple function represents a closed circuit if the relay is operated, while the dashed function represents a closed circuit only if the relay is *not* operated. The conventional symbols for these two types of relay are shown in Fig. 5[3].

Combinations of functions are realized by series and parallel arrangements of switches. Clearly two switches in series represent a closed circuit if and only if each switch is closed, and this is the condition for an "and"

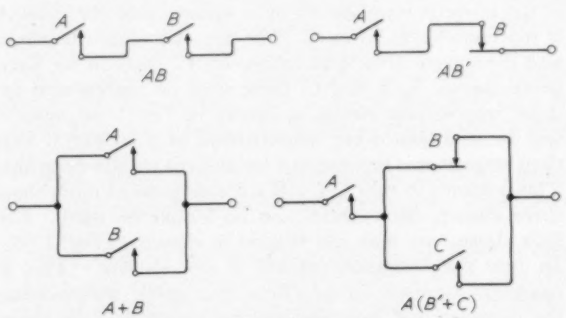


Fig. 6. *Boolean representations of typical relay circuits*

circuit. Similarly, two switches in parallel are a closed circuit if either or both switches are closed, and this is an "inclusive or" circuit. The representations of some typical functions are shown in Fig. 6.

Boolean Algebra can also be used to represent the operation of gating circuits.

Consider the circuit shown in Fig. 7, and suppose that A and B are either at 0 volts or at $-v$ volts. Then if A and B are both at 0 volts, current flows through the diodes and the resistor so that the output terminal C is at approximately 0 volts (neglecting the drop across the diode). If A now changes to $-v$ volts while B remains at 0 volts, current is still taken from B to keep C at 0 volts, and diode A is then back biased so that no conduction takes place through that path. If both A and B are at $-v$ volts, then C must also change to $-v$ volts. Thus this circuit is an "and" circuit, and is represented by the equation

$$C = AB.$$

By a similar argument it will be seen that the circuit in Fig. 8 is an "or" circuit, represented by

$$C = A + B.$$

A "not" function can represent a single-stage amplifier, of the form shown in Fig. 9. In this circuit, if point A is at 0 volts, then the potentiometer chain ensures that the transistor is cut off, and the point B will be at $-V_c$ volts. If point A is at $-v$ volts, then the transistor conducts heavily (in fact it bottoms) and point B is at nearly 0 volts. The output may be clipped at $-v$ volts so as to provide a standard signal level. Clearly this circuit fulfils the required conditions and

$$B = A'.$$

Fig. 10 shows a bistable trigger circuit which is a simple form of memory unit. Hereafter I shall refer to any bistable circuit as a flip-flop: although this usage is not universally adopted in this country, it is a common one and is convenient for our present purposes.

The flip-flop has two stable states: with the left-hand transistor bottomed and the right-hand one cut off, or with the right-hand transistor bottomed and the left-hand one cut off. The collector of the bottomed transistor will be at nearly 0 volts, while the collector of the cut off transistor will be at $-V_c$ volts. These two voltage levels can be taken as representing "0" and "1" respectively, and the state of the flip-flop can be determined by taking outputs from the collectors as shown. If we denote the output voltage from the left-hand collector by Q, then the output voltage from the right-hand collector will represent Q', and Q can then be regarded as the quantity stored in the flip-flop. The state is changed by

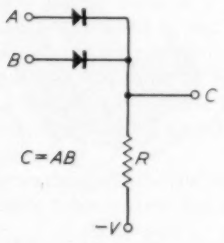


Fig. 7. *Diode "and" gate*

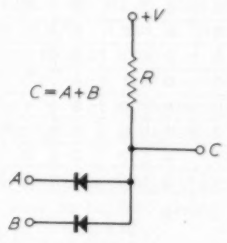


Fig. 8. *Diode "or" gate*

Fig. 9. *(Right) Inverter circuit*

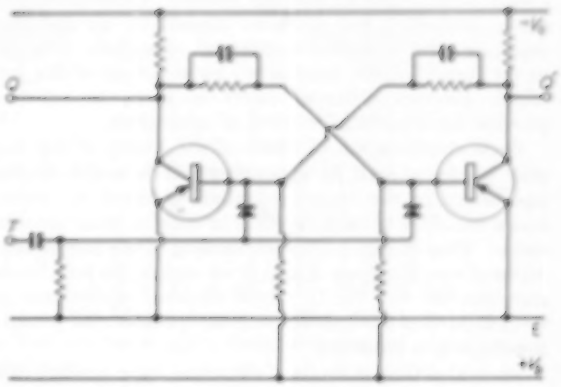


Fig. 10. *Bistable trigger circuit (flip-flop)*

applying pulses to the trigger line T, each incoming pulse causing the state to change.

When we attempt to describe the operation of this circuit in terms of Boolean algebra we encounter a difficulty, because the state the circuit adopts in response to an incoming signal depends on what state it was in before: this means that time has to be introduced into the equation. Now, in a digital computer operations are generally controlled by clock pulses, which appear in a continuous succession on the controlling lines allowing signals to appear at memory unit inputs only at discrete clock pulse times, or "bit times". Memory units are designed in such a way that if they receive an appropriate input at one bit time, they will change to the new state by the next bit time. We will denote the state of any variable X at bit time n by $[X]_n$. We can now consider the operation of the flip-flop, and we see that the input and the state of the flip-flop at time n, $[T]_n$ and $[Q]_n$, together determine the state of the flip-flop at time $n + 1$, $[Q]_{n+1}$. In order to find the law connecting the variables it is helpful to draw up a table, as in Fig. 11. This is called a truth table (another term borrowed from Logic) and displays the value of $[Q]_{n+1}$ which arises from every possible combination of values of $[T]_n$ and $[Q]_n$.

We see from this table that $[Q]_{n+1} = 1$ if

$\{[T]_n = 0$ *and* $[Q]_n = 1\}$ *or* if $\{[T]_n = 1$ *and* $[Q]_n = 0\}$

and we may express this as

$$[Q]_{n+1} = [T']_n[Q]_n + [T]_n[Q']_n$$

or more conveniently

$$[Q]_{n+1} = [T'Q + TQ']_n$$

$[T]_n$	$[Q]_n$	$[Q]_{n+1}$
0	0	0
0	1	1
1	0	1
1	1	0

Fig. 11. *Truth table for flip-flop of Fig. 10*

A	B	NUMBER OF STATE
0	0	0
0	1	1
1	0	2
1	1	3

Fig. 12. *Numbering of states*

Now just as one flip-flop has two distinct states, a group of n flip-flops has 2^n distinct combinations of states, and each one of these (sensed by an *and* gate) could be used to control a different operation. It is easy to see then that the need arises for a circuit of this type which assumes different states in turn, in order to provide for sequential control of operations.

Each combination of states of the group of flip-flops may be designated by a number which is the decimal equivalent of the binary number obtained by writing down the state of each flip-flop in turn in some specified order. Thus in Fig. 12, corresponding to the four possible states of two flip-flops A and B we obtain the four binary numbers 00, 01, 10, 11 whose decimal equivalents are 0, 1, 2, 3. For obvious reasons we speak of this group of flip-flops as a counter.

Several different kinds of electrical logic systems exist, using valves, transistors, magnetic cores, diodes or relays, but they are all essentially the same in that they can be represented by Boolean equations, and so allow a mathematical statement of the required performance of a device to be translated into groups of standard circuits.

We can now see also that the number of circuit components required depends directly on the complexity of the Boolean expression which specifies it, for every extra term means at least an extra diode or transistor, and perhaps a complete extra gate. Furthermore, the inputs to these gates are derived from some other circuit such as a flip-flop, and since this will not stand unlimited loading, extra terms may require emitter-follower stages as well.

Example of Logical Design

To round off this account of the uses of Symbolic Logic in computer work, we will consider the problem of designing a two flip-flop counter, which will count in the sequence 01320 . . . in response to a signal on one of its two input lines, and in the sequence 02130 . . . in response to a signal on the other one. This means that if there is a "1" signal at one input, the two flip-flops must in response to clock pulses take up in turn the combinations of states 00, 01, 11, 10, 00 . . . these being the binary forms of the decimal digits 0, 1, 3, 2, 0. If there is a "1" signal at the other input the flip-flops must take up the combinations of states 00, 10, 01, 11, 00 . . . representing 0, 2, 1, 3, 0 . . . A "1" signal is simply a d.c. voltage which is maintained for as long as the counting sequence is required to continue. When there is no signal on either line, the count is to remain constant, and we are told that there will never be signals on both lines at once.

The counter is to be built of *R-S-T* flip-flops, so called because they have three input lines, R, S and T. A signal on R, the reset line, causes the flip-flop to take up state 0; a signal on S, the set line, causes the flip-flop to take up state 1; and a signal on T, the trigger line, causes the flip-flop to change state. If there is no signal on any line, then the flip-flop remains in the same state, but the response to simultaneous signals on two or more lines is indeterminate.

Consider first the operation of the flip-flop.

We see straight away in this problem how redundancies arise, for plainly we must design the circuit so that those combinations of inputs which give rise to indeterminate outputs are not allowed to occur. That is to say, the conditions RS, ST and RT must be made redundant.

Fig. 13(a) shows the truth table for the flip-flop, omitting all the redundant combinations of inputs. From this we can then write down the full equation for the flip-flop just as we did in the earlier example, by writing down the Boolean sum of all combinations of inputs which give rise to a "1" output. This gives

$$[Q]_{n+1} = [R'S'T'Q + R'S'TQ' + R'ST'Q' + R'ST'Q]_n$$

This is plotted on a Venn diagram in Fig. 13(b), together with the redundant conditions, and from this we can then see that the simplest equations for the flip-flop are

$$\left.\begin{aligned} &[Q]_{n+1} = [S + TQ' + R'T'Q]_n \\ &\text{with } RS = ST = RT = 0 \end{aligned}\right\} \quad \ldots \quad (7)$$

It is interesting to derive this result algebraically, both as an example of the way in which algebraic simplification can be carried out, and as a demonstration of the superiority of the Venn diagram method in cases like these.

We have

$$\begin{aligned} [Q]_{n+1} &= R'S'T'Q + R'S'TQ' + R'ST'Q' + R'ST'Q \\ &= S'(R'T'Q + R'TQ') + R'ST'(Q' + Q) \\ &= S'(R'T'Q + R'TQ') + R'ST' \\ &\qquad \text{since } Q' + Q = 1 \end{aligned}$$

Now since $RS = ST = RT = 0$, we may add as many terms containing these combinations as we like, without affecting the result.

Adding $R'ST$ to $R'ST'$ gives

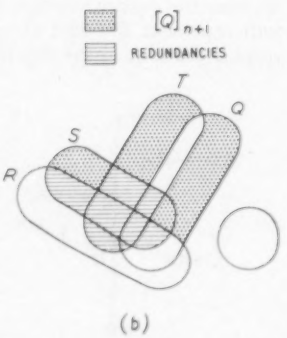
$$R'S(T + T') = R'S$$

and adding then RS we get

$$R'S + RS = (R' + R)S = S$$

$[R]_n$	$[S]_n$	$[T]_n$	$[Q]_n$	$[Q]_{n+1}$
0	0	0	0	0
0	0	0	1	1
0	0	1	0	1
0	0	1	1	0
0	1	0	0	1
0	1	0	1	1
1	0	0	0	0
1	0	0	1	0

(a)

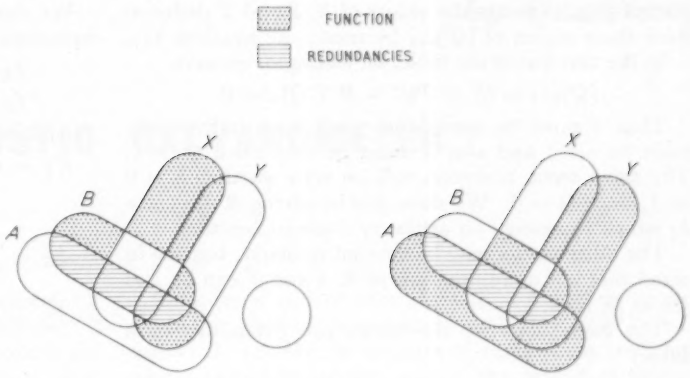


(b)

Fig. 13. *Flip-flop operation; (a) truth table, (b) Venn diagram for* $[Q]_{n+1}$

	$[A]_n$	$[B]_n$	$[X]_n$	$[Y]_n$	$[X]_{n+1}$	$[Y]_{n+1}$
STATIONARY	0	0	0	0	0	0
	0	0	0	1	0	1
	0	0	1	0	1	0
	0	0	1	1	1	1
0213	0	1	0	0	1	0
	0	1	0	1	1	1
	0	1	1	0	0	1
	0	1	1	1	0	0
0132	1	0	0	0	0	1
	1	0	0	1	1	1
	1	0	1	0	0	0
	1	0	1	1	1	0

(a)



(b) (c)

Fig. 14. *Counter operations; (a) truth table, (b) Venn diagram for* $[X]_{n+1}$ *and (c) Venn diagram for* $[Y]_{n+1}$

This gives

$$[Q]_{n+1} = S + S'(R'T'Q + R'TQ')$$

Now, one of the useful identities mentioned earlier was

$$A + A'B = A + B$$

Applying this gives

$$[Q]_{n+1} = S + R'T'Q + R'TQ'$$

We may now add RTQ' and combine it with the last term as before to give the final result.

$$[Q]_{n+1} = [S + TQ' + R'T'Q]_n$$

The algebraic method is much more difficult than the Venn diagram method for seeing the best way of combining terms. In Fig. 13(b), for example, it is immediately obvious that the whole of S can be included in the expression, and it is then a simple matter to find the best way of including the two remaining areas which are not in S.

We now turn to the problem itself, and construct a truth table representing the desired operation of the counter. This is shown in Fig. 14(a). We begin by writing down all possible combinations of the states of the input lines A and B and of the flip-flops X and Y at time n, except that we omit all lines in which A and B are both "1", since in the specification of the problem we are told that the combination AB is redundant.

We can now complete the truth table with the values of X and Y at time $n + 1$ as required by the specification. Thus in the first four lines A and B are both "0", so that the count is to remain stationary, and $[X]_{n+1}$ and $[Y]_{n+1}$ must be the same as $[X]_n$ and $[Y]_n$ respectively. In lines 5 to 8 there is a "1" signal on B, so that the count is to proceed in the order 02130. Thus in line 5 the count at time n is 00 $(= 0)$, so that at time $n + 1$ it is required to become 10 $(= 2)$. In line 6 the count at time n is 01 $(= 1)$ so that at time $n + 1$ it is required to become 11 $(= 3)$. The rest of the table is completed in the same way.

The expressions for the outputs are then obtained from the truth table in just the same way as before, giving

$$[X]_{n+1} = [A'B'XY' + A'B'XY + A'BX'Y + A'BX'Y + AB'X'Y + AB'XY]_n$$

$$[Y]_{n+1} = [A'B'X'Y + A'B'XY + A'BX'Y + A'BXY' + AB'X'Y' + AB'X'Y]_n$$

From the Venn diagrams [Fig. 14(b) and (c)] we see that we can simplify these with the aid of the redundancies to

$$\left.\begin{aligned} [X]_{n+1} &= [(AY + A'B')X + (AY + B)X']_n \\ [Y]_{n+1} &= [(X' + A'B')Y + (AX' + BX)Y']_n \\ &\text{With } AB = 0 \end{aligned}\right\} \quad (8)$$

This is not the simplest possible form, but it is convenient to have the expressions in the form

$$[Q]_{n+1} = [fQ + gQ']_n \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (9)$$

where f and g are Boolean functions of the inputs and outputs.

As a further example of the use of Boolean Algebra (although again it is not to be preferred to the Venn diagram method), we will derive the X expression algebraically.

First we divide the full expression into an X term and and X' term, giving

$$[X]_{n+1} = [(A'B'Y' + A'B'Y + AB'Y)X + (A'BY' + A'BY + AB'Y)X']_n$$

The first two terms in each bracket can be combined, giving

$$[X]_{n+1} = [(A'B' + AB'Y)X + (A'B + AB'Y)X']_n$$

We may now add redundant terms (i.e., terms in AB) where convenient, giving

$$[X]_{n+1} = [(A'B' + AB'Y + ABY)X + (A'B + AB + AB'Y + ABY)X']_n$$

and by combining the terms we have the final result

$$[X]_{n+1} = [(A'B' + AY)X + (B + AY)X']_n$$

By combining equations (7) and (9), we obtain the equations for the inputs to the flip-flops in terms of f and g, since

$$[Q]_{n+1} = [fQ + gQ']_n = [S + TQ' + R'T'Q]_n$$

The left-hand part of Fig. 15 shows the truth table for $[Q]_{n+1}$ obtained from equation (9) for each possible combination of values of f, g, and Q. The right-hand

part of Fig. 15 shows the values of R, S and T deduced from these values of $[Q]_{n+1}$ by means of equation (7).

In the first line of the table, for example, we have

$$[Q]_{n+1} = [S + TQ' + R'T'Q]_n = 0$$

Thus S must be zero (since each term individually must be zero) and also T must be zero since $Q' = 1$. The third term, however, will be zero whether R is 0 or 1, since $Q = 0$. We show this by giving R the value k_1 which represents an arbitrary Boolean constant.

The other values may be derived similarly, bearing in mind that not more than one of R, S and T can be 1 at the same time.

The expressions for the inputs are (from the truth table)

$$R = k_1 f'g'Q' + k_2 f'g'Q + k_4 f'gQ + k_5 fg'Q'$$
$$S = k_3 f'gQ' + k_6 fg'Q + k_7 fgQ' + k_8 fgQ$$
$$T = k_2' f'g'Q + k_3' f'gQ' + k_4' f'gQ + k_7' fgQ'$$

There are obviously several ways of choosing the constants to give simplified expressions for these inputs: one choice is to make each expression independent of Q by putting

$$k_1 = k_2 = k_7 = k_8 = 1$$
$$k_3 = k_4 = k_5 = k_6 = 0$$

and this gives

$$R = f'g', \quad S = fg, \quad T = f'g$$

$[f]_n$	$[g]_n$	$[Q]_n$	$[Q]_{n+1} = [fQ + gQ']_n = [S + TQ' + R'TQ]_n$	$[R]_n$	$[S]_n$	$[T]_n$
0	0	0	0	k_1	0	0
0	0	1	0	k_2	0	k_2'
0	1	0	1	0	k_3	k_3'
0	1	1	0	k_4	0	k_4'
1	0	0	0	k_5	0	0
1	0	1	1	0	k_6	0
1	1	0	1	0	k_7	k_7'
1	1	1	1	0	k_8	0

Fig. 15. (*Above*) *Input equations truth table*

We can now complete the design, for comparing equations (8) and (9), we see that

$$f_X = AY + A'B' \qquad g_X = AY + B$$
$$f_Y = X' + A'B' \qquad g_Y = AX' + BX$$

Therefore

$$\begin{aligned} R_X = f_X'g_X' &= (AY + A'B')'(AY + B)' \\ &= (A' + Y')\,(A + B)\,(A' + Y')B' \\ &= AY' \\ S_X = f_Xg_X &= (AY + A'B')\,(AY + B) \\ &= AY \\ T_X = f_X'g_X &= (AY + A'B')'(AY + B) \\ &= (A' + Y')\,(A + B)\,(AY + B) \\ &= B \end{aligned}$$

and similarly

$$R_Y = AX, \quad S_Y = AX', \quad T_Y = BX$$

In the simplification of these expressions we make use of all the usual properties of Boolean functions, together with the redundant condition

$$AB = 0$$

From these equations, we can now draw the circuit diagram, and this is shown in Fig. 16. All the resistors go to the $-V$ supply, as in Fig. 7.

Notice that clock pulses are necessary on all inputs: in this particular example though we are able to economise on diodes by a judicious grouping of the inputs, applying the clock pulses to the two gates M and N only.

I will not claim that this method of solution is realistic in this particular case, because the problem is so artificially simple, but it does serve to demonstrate the sorts of ways in which the principles of Symbolic Logic can be applied in computer design.

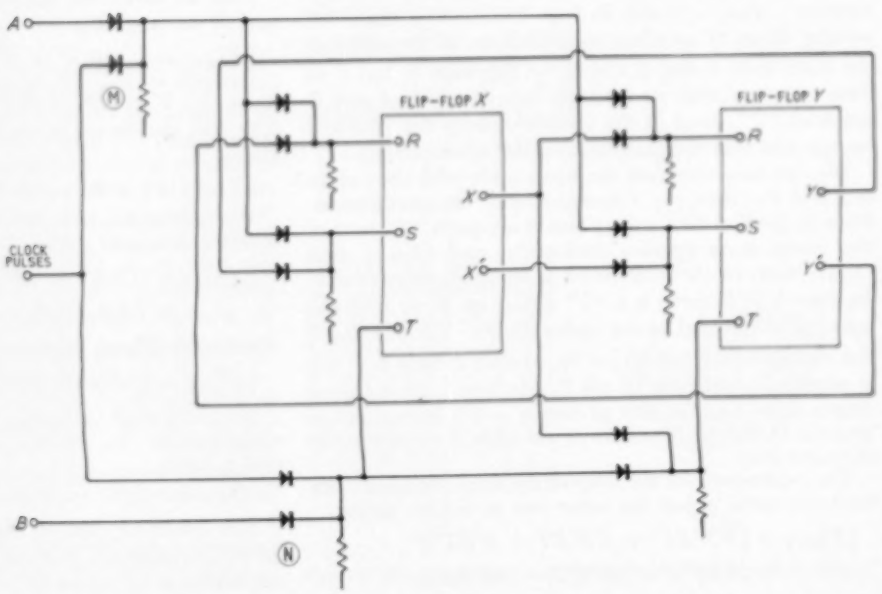


Fig. 16. (*Right*) *Circuit diagram of counter*

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THE BALLISTIC GALVANOMETER

The questions, "What is a ballistic galvanometer?" and "When is a non-ballistic galvanometer ballistic?" are old ones which I expect you will find thrashed out in the traditional practical textbooks like Watson. But, nowadays, they take on a new importance in elementary teaching, since the approach in rationalized m.k.s. units (see, for example, "The Teaching of Electricity", published by John Murray for the Science Masters' Association) requires the use of a ballistic galvanometer as a fluxmeter. The late F. A. Meier, who contributed many of the experiments in this book, was emphatic that you don't need what is officially rated as a ballistic galvanometer in order to be able to do flux-change experiments. Ordinary sensitive current meters, providing they satisfy certain requirements, work reasonably well. The chief requisite, of course, is that they shall not be deliberately damped by having the coil wound on a conducting former. That kind is out of the picture for good. It might be worth while to examine what "reasonably well" means, and to see how an ordinary commercial instrument (which may be recognizable but which shall be anonymous) features in this respect.

Ballistic Use

I used to think that a ballistic galvanometer had to satisfy three conditions—a long period of swing, very low damping, and a high resistance. Of these, the first is desirable, the second essential, and the third, it turns out, rather a nuisance if you can get adequate sensitivity without it.

In ballistic use, the charge q to be measured passes through the coil in a time which is short compared with the period T of free swing, and must always be considerably less than $T/4$. The suspended system experiences an angular impulse $\int C dt$ which, since the torque at each instant is proportional to the current i, is proportional to $\int i dt$, that is, to q. The system then swings out until the kinetic energy has *mostly* gone in work done in twisting the suspension, and the resulting maximum deflection is the 'throw' θ. The "Concise Oxford Dictionary" says, incidentally, that "ballistic" comes from the Greek word for a throw; and the galvanometer appears to be one of the few cases where it is correctly used. It will not perhaps influence the cause of disarmament to observe primly that the last effective ballistic missile (if we except the domestic rolling pin) was the javelin, nor clarify the recent problems of cricket to remark that if a ball is ballistic it is a no-ball. But, as long as we remember that it is the *throw* and not the static deflection that matters in a ballistic meter, that will suffice; there's enough pedantry about already, anyhow, most of it masquerading as literacy.

Consider a coil of area A, with n turns, resistance (including that of its suspension) G, in a radial flux-density B. Let I be the moment of inertia, c the restoring torque per radian twist, and T the period of swing. Write the product BAn as Φ (the flux linkage) for short.

Clearing off the static condition first, a steady current i_0 gives a static deflection θ_0, where $\Phi i_0 = c\theta_0$; and calling the static ratio i_0/θ_0 's', we have $c = \Phi s$. The period of swing T, which will also be written as usual in the form $T = 2\pi/\omega$, is given by $T = 2\pi\sqrt{I/c}$. So, substituting for c

$$\omega^2 = \Phi s/I \qquad \ldots \quad (1)$$

Now, let the coil be swinging at any instant with angular velocity $d\theta/dt$, then the induced e.m.f. is $\Phi d\theta/dt$. If the resistance in the external circuit is R, the current passing is $i = \Phi(d\theta/dt)/(R + G)$. The torque opposing the motion of the coil is Φi, which is $\Phi^2(d\theta/dt)/(R + G)$. In the expression for the motion of the coil, this has to be divided by the moment of inertia I to get the factor usually written as $2k$, which is the resisting couple per unit angular velocity, the torque when $d\theta/dt = 1$. So

$$2k = \Phi^2/(R + G)\, I \qquad \ldots \quad (2)$$

Now, in the equation of motion of the damped swinging of the coil

$$d^2\theta/dt^2 + 2k d\theta/dt + \omega^2\theta = 0$$

with the usual notation, the solution is

$$\theta = \theta_0 e^{-kt} \cos\{\sqrt{(\omega^2 - k^2)}t\},$$

where $\theta = \theta_0$ when $t = 0$. Oscillations occur if $\omega^2 > k^2$, and the critical case is when $\omega^2 = k^2$, or $k^2/\omega^2 = 1$.

Now, from (1) and (2)

$$\frac{k^2}{\omega^2} = \frac{\Phi^2}{4(R + G)^2\, Is}$$

If D be the initial-damping external resistance that makes $k^2/\omega^2 = 1$, then $\dfrac{\Phi^2}{4Is} = (D + G)^2$, whence $\dfrac{k^2}{\omega^2} = \dfrac{(D + G)^2}{(R + G)^2}$. Taking square roots, and remembering that $\omega = 2\pi/T$,

$$k = \frac{2\pi}{T}\,\frac{(D + G)}{(R + G)}, \text{ so}$$

$$k = \frac{2\pi}{T}\,\frac{(1 + D/G)}{(1 + R/G)} \qquad \ldots \quad (3)$$

We have made one rash assumption here—that *all* the damping is electromagnetic. Really k in the differential equation is the sum of two terms, an electromagnetic k_e and a mechanical k_m; equation (3) refers to k_e.

Compromise

If the meter is to be useful ballistically, k should be less than about 0·2, when the damping is less than 10 per cent of the corrected throw. If it is as high as 0·5 then you are running into trouble, though a damping correction as high as 20% could be tolerated, I suppose.

It is readily possible to determine the mechanical damping experimentally by observing the open-circuit swing. In the particular meter studied, this was approximately 0·1, which leaves only a further 0·1 for k_e, instead of 0·2, available as the tolerable electromagnetic contribution. So we will see what use can be made of it within these limits.

Given $T = 2$ sec, $D = 100\ \Omega$, $G = 25\ \Omega$,

$$k = 5\pi/(1 + R/25) \quad .. \quad .. \quad .. \quad .. \quad (4)$$

so that, if k is to be less than 0·1, R is about 3 800 Ω.

Where does the compromise come in? Well, a simple example on the earth inductor will show. Since the charge circulating is given by (flux change)/$(R + G)$ in any operation, if the meter gives a full-scale throw for about 10^{-6} coulomb, then the sort of flux change that you have to aim at is about 4×10^{-3} volt-second. This means that (taking about 4×10^{-5} volt-sec/sq.m as a rough value for the flux density of the earth's field) your earth-inductor must have area-turns of about 100 sq. m. If you are dealing with smaller flux-densities, of course you have to choose between sensitivity and compliance with ballistic conditions.

To sum up, then, while it is of course true that the ideal ballistic conditions are served by having a small k which goes with a large T, yet even if T is relatively short, we can obtain ballistic conditions provided (1) the makers give us a small D/G, and (2) we then use it in circuits with a large R/G.

SPUTTERED-DIELECTRIC CAPACITORS

Among the many types of capacitors being developed by the Capacitor Division of Standard Telephones and Cables is the sputtered-dielectric type. This is being investigated with a view to producing miniature low-voltage capacitors having characteristics somewhat similar to those of mica capacitors, but with improved capacitance : physical size ratios.

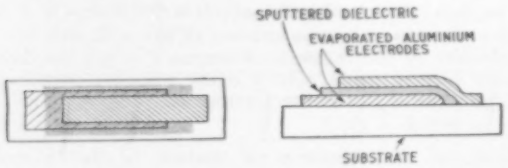

Fig. 1 (*Above*). *This drawing indicates the method of formation of a sputtered-dielectric capacitor*

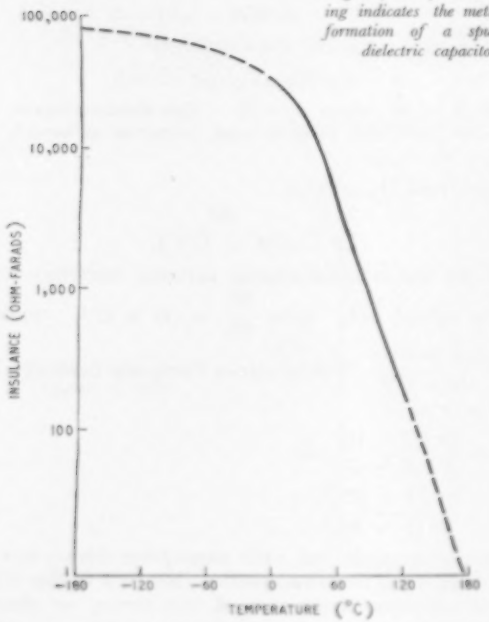

Fig. 2. *The curve insulance vs temperature for a sputtered silicon oxide dielectric*

These capacitors are formed by first evaporating an aluminium electrode on to a substrate such as glass, then depositing a film of metal oxide dielectric by reactive sputtering and finally evaporating on a top electrode of aluminium (Fig. 1). Dielectric films of silicon oxide, tantalum oxide, niobium oxide, zirconium oxide and titanium oxide have been produced by sputtering.

S.T.C. are proposing to commence production with the introduction of a sputtered-dielectric capacitor for 6-V operation. This will be initially about 1,000 pF and when protected in resin to give a H1 performance the dimensions will be 0·3 in × 0·25 in × 0·125 in. With further development and by the use of multi-layer techniques it is expected that it will be possible to increase the capacitance without increasing the overall size; the size being determined almost entirely by the thickness of the substrate and the protective medium.

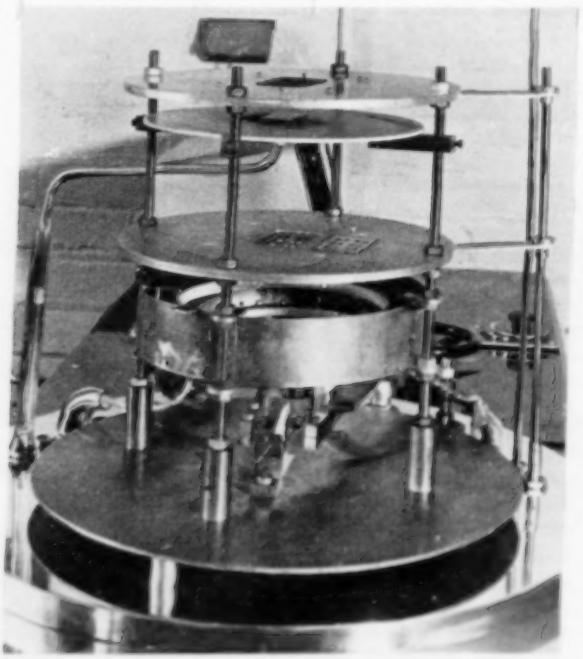

Fig. 3. *An illustration showing the general disposition of the aluminium source and silicon cathode within a vacuum chamber used for electrode evaporation and dielectric sputtering*

Tunable Rejection Filters

By L. A. Moxon*, B.Sc., A.M.I.E.E.

Optimum design for tunable rejection filters is a problem which has been studied with the main object of preventing interference to receivers from a number of adjacent transmitters. This study has led to firm conclusions regarding right and wrong ways of designing this type of filter; these conclusions are unrelated to current design practice and the author's approach would therefore appear to provide a new angle on this problem.

In some applications the complication of having more than one tuning adjustment may be inadmissible, and it is convenient to consider first what is the best that can be done in these circumstances, and secondly, the best way of using additional controls when these are acceptable; in so doing, important considerations are the depth and width of the rejection notch, the insertion loss, and the extent to which performance is affected by component tolerances.

Tunable rejection filters usually employ one or more resonant circuits but, in principle, any of the numerous forms of bridge circuit can be used, and one could also adopt a classical filter-theory approach. The latter can be regarded as a method of combining a number of resonant circuits but is difficult to apply to tunable filters as it usually requires ganging or individual adjustment of dissimilar elements. Whether the basic element is a tuned circuit or a frequency-sensitive bridge network, rejection is essentially a balancing process involving, in general, both reactance and resistance. Reactance is essential since the balance is required to be frequency dependent, whereas resistance arises from imperfections of components and may or may not be important, depending on the degree of rejection required and the way in which the circuit is arranged. There are a number of ways of achieving a null by balancing resistance as well as reactance and, in principle, the resistance balance need not be frequency sensitive, so that a single preset adjustment of resistance should provide additional rejection without adding to the number of tuning controls. In practice, however, when covering an appreciable tuning range, it is only possible to maintain an approximate balance. It will be shown that circuit arrangements tend to fall into two categories, those which minimize the role of resistance so that an approximate balance is often acceptable, and those which maximize the dependence of rejection ratio on resistance so that performance is unduly dependent on frequency as well as on component tolerances. In the first category we find the bridged-T filter, with the conventional bifilar-T in the second category but transferable to the first by a simple modification. Basically, however, the depth and width of the rejection notch, the bandwidth of the signal to be rejected, and the additional rejection obtainable by the cascading of ganged filter sections are linked by single relationships which are largely independent of the circuit technique used.

* Royal Naval Scientific Service.

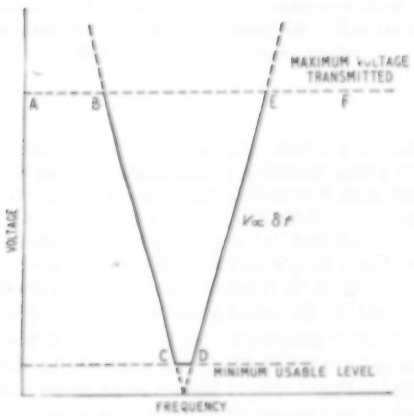

Fig. 1. *Response of an idealized single-section filter*

Basic Principles

Suppose we have two quantities either or both of which are functions of ω, the difference between them, θ, being equated to zero at frequency ω_0; i.e.

$\theta = A F_1(\omega_0 + \delta\omega) - B F_2(\omega_0 + \delta\omega)$

where $A F_1(\omega_0) = B F_2(\omega_0)$

Assuming $\delta\omega \ll \omega_0$ and $F_1 \neq F_2$ we have (using Taylor's Theorem)

$$\theta \simeq \delta\omega[A F_1'(\omega) - B F_2'(\omega)] \quad \ldots \quad \ldots \quad \ldots \quad (1)$$

θ can be, for example, the impedance of an inductance and capacitance in series, being then equal approximately to $\delta\omega\, 2L$. In general, the balancing of reactances can be associated with the balancing of the voltages across them or currents through them so that Equ. (1) also represents the amplitude response of the network and produces a rejection curve which is basically a V-notch as illustrated in Fig. 1, the voltage being proportional to the tuning-error $\delta\omega$ provided $\delta\omega/\omega$ is not too large. Superimposed on this, however, we have a lower limit set by imperfections of balancing or the finite bandwidth of the signal to be rejected and an upper limit set by the fact that, for

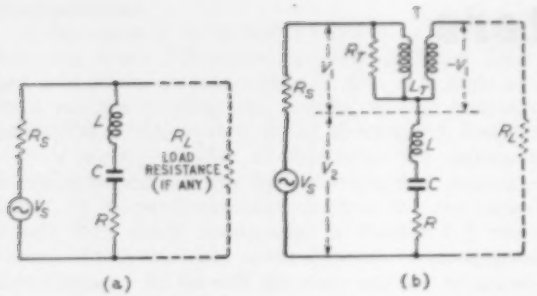

Fig. 2. *A series-tuned filter: (a) shows the basic circuit, while (b) shows the basic circuit with a transformer and resistance to balance the residual voltage; balance occurs when* $V_1 = V_2$

one reason or another, the output voltage cannot reach infinity. These limits are indicated in Fig. 1 by the horizontal dotted lines, so that with suitable scales and some intuitive rounding of corners A B C D E F might be expected to represent, approximately, the response of any single-section rejection-filter.

Fig. 2(a) shows a conventional series-type rejection-filter for which the minimum response is determined by the finite resistance of the coil being given approximately by $V_S R/R'$; where R' is equal to $R_S R_L/(R_S + R_L)$. The maximum response, less 3 dB, occurs when the reactance of the series circuit is equal to the resistance of the source and load in parallel; i.e., when $R' \simeq 2\ (\delta\omega)L$. Defining the distance R'/L between these points as the notch width, W_n, the voltage response at the rejection frequency is inversely proportional to $Q\ W_n/\omega$. This is not a basic limitation, however, since the residual voltage can be balanced out as shown in Fig. 2(b). If T is an ideal phase-reversing transformer, of unity voltage ratio, the voltages will be as shown in the diagram giving infinite rejection when $V_1 = V_2$; i.e., when $\omega L = 1/\omega C$ and $R_T = R$. As shown below, this argument is readily extended to cover the practical case of a 'non-ideal' transformer.

From inspection of Fig. 1, the voltage rejection ratio which can be exceeded over any finite bandwidth W_r is equal to W_n/W_r where W_n is the notch width. The circuit of Fig. 2(b) enables this limit to be reached by a combination of two processes, the balancing of inductive against capacitive reactance (i.e., tuning) and the balancing of the resistances. In principle the task can be shared out as desired between the two mechanisms but it will be observed that the need for tuning is inherent in the frequency dependence of the system whereas the resistance balance need only be invoked if $W_n/W_r > QW_n/\omega$. A limit to the balance achievable is set by component tolerances and the variation with frequency of the effective values of resistance so that the achievement of large rejection ratios is assisted by making Q as large as possible, thus minimizing the reliance which must be placed on the resistance balancing.

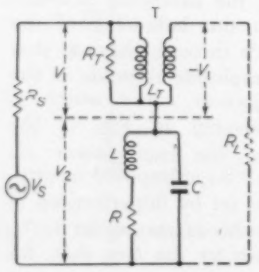

Fig. 3. *A conventional form of a bifilar-T circuit. Voltage designations compare with* Fig. 2 *(b); balance occurs when* $V_1 = V_2$

Assuming that a tunable filter of this type has been designed to give the best performance practicable with a single variable parameter, and if the rejection is still inadequate, further improvement can be obtained as follows:

(a) If the rejection-ratio falls short of W_n/W_r, where W_r is the bandwidth of the signal to be rejected, the amount by which it does so can be exploited by giving the user a second knob for adjusting the resistance balance.

(b) By cascading filters. If the sections are ganged then, so far as the user is concerned, there is still only a 'single variable' but the ganging tolerance $\pm\ \delta\omega_G$ makes the effective useful length of the intercept CD (Fig. 1) for the added circuits equal to $2\ \delta\omega_G$; it is therefore not possible to rely on an additional rejection greater than $W_n/2\delta\omega_G$. Since ganging tolerances tend to be of the same order as circuit bandwidths, the maximum rejection ratio will not be reliably increased by applying resistance-balancing techniques to more than one out of two or more ganged sections.

Method (a) is considered to have relatively limited application since, if two knobs are permissible, it is obviously preferable to use two sections each with an approximate resistance balance. This not only provides, as a rule, greater attenuation, but results in less critical tuning since for a given total attenuation the bandwidth is greater by the square root of the attenuation ratio, as is obvious from inspection of Fig. 1.

It is not sound practice to cascade filters of the type shown in Fig. 2, since one filter provides a low value of R_S for the next one and means that the added rejection is obtained by balancing alone with little or no contribution from the shunting effect on the signal line. This difficulty is easily avoided, as discussed later, by alternating sections of the Fig. 2 type with bridged-T sections (Fig. 4).

It is possible to rely on balancing for the whole of the attainable rejection ratio W_n/W_r. The above argument has shown that this is bad engineering, yet it is common practice to go even further in the wrong direction by using the tuned circuit to enhance the voltage which has to be disposed of by balancing, as in the circuit of Fig. 3. This is the same as Fig. 2(b) with the series tuned circuit replaced by a parallel one, and will be recognized as the conventional form of bifilar-T rejection-filter. At balance the mechanism is identical in the two cases with R replaced by Q^2R in the case of Fig. 3. A realistic treatment requires to take into account the inductance L_T of the transformer and this has been done for the conventional bifilar-T circuit using a series of star-delta transformations to derive the balance conditions and other properties of

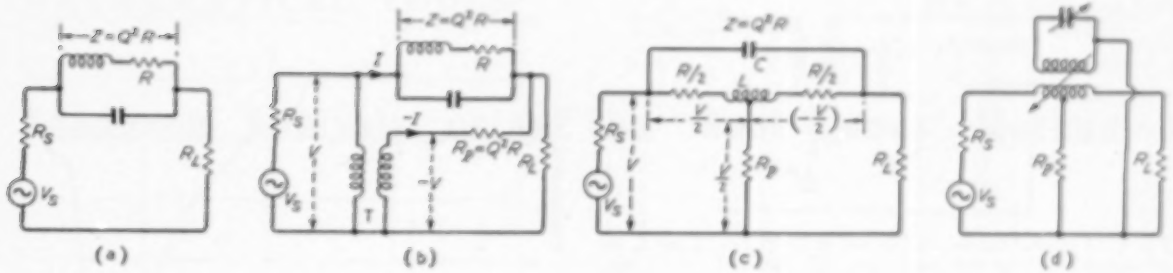

Fig. 4. Parallel-tuned rejection filters: (a) shows a simple parallel-tuned trap in which $Z \gg R_S R_L/(R_S+R_L)$ at resonance; (b) is a bifilar resistance-balanced version of (a) with T as a phase-reversing transformer; (c) is a bridged-T resistance balanced version of (a), and (d) is a modified form of (c) with one side of the tuning capacitor earthed

the circuit[1]. This can be readily extended to the series circuit, but if our interest is only in the balance condition, this can be written down by inspection of Fig. 2(b) or Fig. 3; the requirement is that the impedance of the resonant circuit shall equal that of R_T, ωL_T in parallel so that replacing this by an equivalent series impedance $R_S + j\omega L_S$ we have

$$\left.\begin{aligned} R &= R_S \\ \omega L_S = \omega L - \frac{1}{\omega C} &\text{ ; i.e., } 2\frac{\delta\omega}{\omega} \mathrel{\hat{=}} L_S/L \end{aligned}\right\} \quad .. \quad (2)$$

where $\delta\omega/\omega$ is the fractional detuning required.

To make use of this result, we have to substitute for R_S, L_S

$$\left.\begin{aligned} R_T\,\omega^2 L_T^2/(R_T^2 + \omega^2 L_T^2) &= R_S \\ L_T\,R_T^2/(R_T^2 + \omega^2 L_T^2) &= L_S \end{aligned}\right\} \quad .. \quad .. \quad (3)$$

From the first parts of Equs (3) and (2) we have for any given transformer and tuned circuit

$$R_T = \omega^2 L_T^2/2R \pm \sqrt{\omega^4 L_T^4/4R^2 - \omega^2 L_T^2} \quad .. \quad (4)$$

For a real solution, therefore, we require $\omega L_T > 2R$. At this lower limit for L_T we have $R_T = \omega L_T$ and $\omega L_S = R_S = R$ so that from Equ. (1), $\delta\omega/\omega = R/2\omega L = 1/2Q$ and at balance the circuit is detuned on the inductive side of resonance by half the '3 dB bandwidth' B. The line voltage, however, will be a minimum with similar detuning at the other side of resonance; i.e., when $2\omega L_T \mathrel{\hat{=}} -(\omega L - 1/\omega C)$. The output, therefore, rises less rapidly with capacitive detuning and the response is assymetrical to the extent of 3 dB at frequencies $\pm B/2$ from the natural resonance. The assymetry decreases with further detuning and with increase of L_T.

It has been shown that the rejection process can be regarded as involving either or both of two mechanisms, and that best and worst systems are characterized by whether they maximize the role of one mechanism or the other. One 'best' system has been studied in detail, but since there are basically two ways (i.e., series and parallel circuits) of using a tuned circuit as a rejection filter, one might expect to find at least two equally good systems. Starting with Fig. 4 (a), we have the problem of balancing out a residual voltage given approximately by $V_S R_L/(Q^2 R)$, and this could be achieved by the arrangement of Fig. 4 (b) with $Q^2 R = R_p$ as the obvious balance condition. Alternatively, we can employ the circuit of Fig. 4 (c) which will be recognized as the well-known bridged-T filter and has been fully analysed in existing literature[2]. This latter system has the advantage of evading the problems of leakage reactance and stray capacitance which exist in the case of the bifilar windings; however, except near resonance, the main consequence of stray capacitances is slightly to modify the effective value of inductance, whereas leakage reactance mainly affects the transformer ratio and can be compensated in obvious ways. A practical disadvantage of the conventional bridged-T filter is the need to insulate both ends of the tuning capacitor. The alternative, Fig. 4 (b), suffers from the more serious disadvantage of appreciable insertion loss unless ωL_T is large compared with $R_L R_S/(R_L + R_S)$; if this demand is met, the self-capacitance of the bifilar windings may prove troublesome. Fig. 4 (d) indicates one way of circumventing the capacitor insulation problem; this is mainly of importance for ganged circuits however, and there is no real problem when the circuits are separately tunable. It will be noticed that, by providing a high source impedance near resonance, a bridged-T section assists the operation of a subsequent bifilar circuit of the Fig. 2 (b) type, whereas the latter provides the low input impedance desirable for a subsequent bridged-T filter, so that a multi-section filter should employ these types alternately.

For the bridged-T configuration, the notch width between '3 dB down' points is obtained by equating the effective circuit reactance to $R_S + R_L$ [i.e., $W_n L = Q^2 R^2/(R_S + R_L)$] and the balance condition from inspection of Fig. 4 (c) is $R_b = Q^2 R/4 = Q\omega L/4 = (Q/4)/(L/C)$. With optimum coil design Q tends to be independent of ω so that if the resistance is to be balanced over a range of frequencies, it is necessary either to depart from optimum design or vary L and C simultaneously. With variation of L or C at constant Q the percentage of resistance unbalance is equal to the percentage frequency shift so that the resistance contributes only about 14 dB of rejection at the edges of a 1·5 : 1 tuning range. The situation with the arrangement of Fig. 2 (b) is a little different; referring to Equs (2) and (3) we require $R_T\omega^2 L_T^2/(R_T^2 + \omega^2 L_T^2) = K\omega$ where R has been replaced by $K\omega$ to give effect

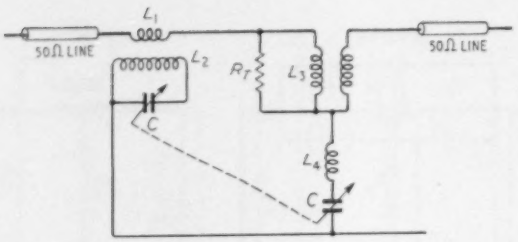

Fig. 5. *A practical rejection filter for 6–8 Mc/s.* [L_1, L_2, L_3 *and* L_4 *are all wound on Neosid 900 grade cores, type 10D.* $L_1 = 1$ *turn,* $L_2 = L_4 = 21\ \mu H$ *(21 turns),* $L_3 = 1\ \mu H$ *(4 turns), and* $R_T = 5\ \Omega$]

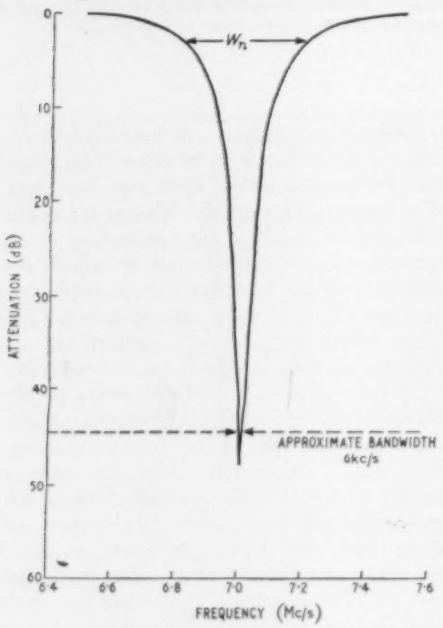

Fig. 6. *A typical measured response of the filter given in* Fig. 5. ($W_n \simeq 5\%$ *of the mean frequency*)

to the normal tendency for $Q(=\omega L/R)$ to remain constant with frequency. Equating $\frac{d}{d\omega}\frac{\omega}{R_T^2 + \omega^2 L_T^2}$ to zero, we have $R_T = \omega L_T$, so that the lowest possible value of L_T is also the one which, in general, gives the minimum frequency dependence of the resistance balance.

Insertion loss at frequencies remote from resonance tends to be very small with both the bridged-T and the bifilar-T filters, being due mainly to the balancing resistance which is respectively in parallel and in series with the source of signal so that the fractional loss of available power is given approximately by R_s/R_b or $4R_T/R_S$ as appropriate.

Practical Results

The foregoing considerations resulted in the design of a number of filters tunable over roughly half an octave and together covering most of the h.f. band. Fig. 5 shows the circuit arrangement, C being a two-gang capacitor; this is a two-section filter using one parallel-tuned and one series-tuned section, resistance balance

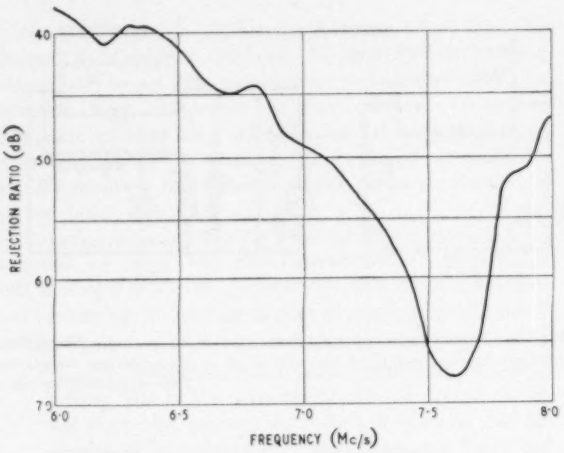

Fig. 7. *The measured rejection ratio vs frequency for the filter given in* Fig. 5. *The rejection at* $\pm$ 3 *kc/s is roughly constant in the region of* 40 *to* 50 *dB*

being applied to the series-tuned section only, since it was argued that ganging tolerances would not justify applying it to both. This point is debatable, however, since the two resistance balances could have been pre-set at different frequencies thereby increasing the average rejection ratio. It should be noted that one function of CL_1L_2 is to prevent the flow, at resonance, of large currents which could cause physical damage to L_3L_4.

Fig. 6 shows a typical response curve, and Fig. 7 shows an example of the rejection as a function of frequency with the resistance balance fixed.

Conclusion

Experimental results are qualitatively in good accord with the simplified theory of rejection filters outlined above, and it would appear that the design of such filters can, in general, be considerably improved.

The filter of Fig. 5 is not of course the 'last word' in design; its performance could be surpassed by, for example, cascading a sufficient number of elementary series and parallel-tuned circuits, and such an arrangement would have the advantage of providing a more rectangular notch. It will be noted, however, that such a system could always be further improved by the addition of resistance balance to one or more sections in accordance with Figs. 2 (b) or 4 (d).

The argument has been based mainly on typical LC filters but is applicable in part to bridge-type filters including Wien bridge, Twin-T, and other types of RC filter. These have low effective Q-values, and, since they rely mainly on balance, tend to be non-optimum.

Acknowledgement

The author is indebted to Mr. T. Searle for the experimental results.

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Feedback Analysis using the 'Root Locus' Method—1

Suppose that we have a negative-feedback amplifier in which the voltage gain of the amplifier is A and that of the feedback path is β (usually the effect of the feedback path does not depend appreciably on frequency, and β can be taken as a number between 0 and 1). Then it is well known that the open-loop gain is $A\beta$ and the closed-loop gain A_f is given by

$$A_f = A/(1 + A\beta) \quad .. \quad .. \quad .. \quad .. \quad (1)$$

Now the gain A depends upon the nature of the input —in particular, for a sinusoidal input A depends upon the frequency. A can be expressed in terms of the Heaviside operator p or the Laplace-transform quantity s, and formally A is the same in either case. We shall work in terms of p, as in previous 'Tools', so that the steady-state response at frequency $\omega/2\pi$ is given by replacing p by $j\omega$, but this is irrelevant here. If we need to consider the response to any particular input as a function of time, we can obtain this from knowledge of A (as a function of p) and of the input as explained in previous 'Tools', but here again this is not the point. Our present objective is to find out all we can about the essential features of the feedback system (stability, damping, etc.) without having to go to the trouble of a complete determination of the response as a function of time. For the relevant features of that response become obvious if we have sufficient knowledge about the behaviour of the expression $(1 + A\beta)$, as will be explained.

For most practical amplifiers, A is a number (say A_0, the low-frequency gain) multiplied by the quotient of two polynomials in p, say $N(p)$ and $D(p)$. If $A_0\beta$ is called K for short, we can thus write

$$A\beta = K\frac{N(p)}{D(p)} \quad .. \quad .. \quad .. \quad .. \quad (2)$$

For a three-stage RC-coupled amplifier with valves of mutual conductances g_1, g_2, g_3, coupling resistances R_1, R_2, R_3 and shunt capacitances C_1, C_2, C_3 respectively, the gain of each stage is of the form $gR/(1 + pCR)$, and thus, if $T_1 = C_1R_1$, $T_2 = C_2R_2$ and $T_3 = C_3R_3$ the overall gain becomes

$$A = \frac{g_1R_1g_2R_2g_3R_3}{(1 + pT_1)(1 + pT_2)(1 + pT_3)} \quad .. \quad (3)$$

and this is in the form of Equ. (2) if

$$\left.\begin{aligned} K &= g_1R_1g_2R_2g_3R_3\beta/T_1T_2T_3 \\ N(p) &= 1 \\ D(p) &= (p + 1/T_1)(p + 1/T_2)(p + 1/T_3) \end{aligned}\right\} \quad .. \quad (4)$$

This example is included only to explain the significance of the expressions $N(p)$ and $D(p)$, which are usually easy to determine from the circuit given, as in this case. The root-locus method is a means of finding out the required information about the behaviour of $(1 + A\beta)$ when we know $N(p)$ and $D(p)$ and K varies. In most practical cases, intervalve couplings are individually of a fairly simple nature, so that the terms in $D(p)$ and $N(p)$ which represent them are also simple, usually linear or at worst quadratic. Hence $N(p)$ and $D(p)$ usually arise in the form of products of quite simple factors, as in Equ. (4).

For clarity, we shall follow the usual practice and distinguish between

poles: numbers which can be substituted for p to make $D(p)$ zero and therefore to make $A\beta$ infinite.

zeros: numbers which can be substituted for p to make $N(p)$ zero and therefore to make $A\beta$ zero.

roots: numbers which can be substituted for p to make $(1 + A\beta)$ zero.

Now the essential information about a feedback amplifier is easily obtained if we know the roots. For stability, for example, no root must have a positive real part. The basic problem is therefore to determine the roots when the poles, the zeros and the quantity we have called K (low-frequency open-loop gain in the RC-coupled 3-stage amplifier case discussed above) are known. In simple cases it is quite easy to find the roots algebraically, but this is difficult if $N(p)$ and/or $D(p)$ are of high degree. The so-called root-locus method (References 1-4) is one that enables an approximate solution to be obtained graphically, and that is often sufficient.

A root occurs when $A\beta$ has the value -1, that is to say, the magnitude of $A\beta$ is unity and the associated phase angle is 180°. The magnitude depends very much upon K, but the phase angle does not. Essentially, the root-locus method consists in finding all the complex numbers $x + jw$ which can be substituted for p to make the phase-angle of $(1 + A\beta)$ equal to 180° and plotting them, without in the first instance worrying about the value of K for which they are roots. In practice, we may only need to determine a few roots in strategic areas, and our problem will then be solved adequately by graphical interpolation between these. For example, if we want to know the condition for stability, we may soon find that only one branch of the root-locus is in any danger of crossing the imaginary or ω-axis, and we can devote our main attention to plotting that branch in the immediate neighbourhood of the crossing-point. Only

after the relevant part of the root-locus is determined need we find the value of K which makes $A\beta$ of unit magnitude for any particular point of it.

Consider as an example the case

$$A\beta = K\frac{p+2}{(p+4)(p+3+j5)(p+3-j5)} \qquad (5)$$

for which, in the notation of Equ. (2),

$$\left.\begin{aligned} N(p) &= p+2 \\ D(p) &= (p+4)(p+3+j5)(p+3-j5) \end{aligned}\right\} \qquad (6)$$

so that there is one zero $-2+j0$, and three poles $-4+j0$ and $-3\pm j5$. These are first plotted as in Fig. 1; it is customary to indicate a pole by a cross on the point and a zero by a circle round it. The same scale must be used along the horizontal or α-axis as up and down the vertical or ω-axis. Now let X or $\alpha+j\omega$ be any other point, so that in Fig. 1 α is negative* and ω positive. Then if X is joined to Z, and $XZ = r$ and $\angle OZX = \phi$, we have

$$\alpha-(-2)+j\omega = r\exp(j\phi) \qquad (7)$$

If X is also joined to the poles, and these are at distances R_1, R_2, R_3 and in directions determined by θ_1, θ_2, and θ_3 as indicated in Fig. 1, we also have

$$\left.\begin{aligned} \alpha-(-5)+j\omega &= R_1\exp(j\theta_1) \\ \alpha-(-3)+j(\omega-5) &= R_2\exp(j\theta_2) \\ \alpha-(-3)+j[\omega-(-5)] &= R_3\exp(j\theta_3) \end{aligned}\right\} \qquad (8)$$

and therefore, if $(\alpha+j\omega)$ is substituted for p in Equ. (5),

$$\left.\begin{aligned} \text{magnitude of } A\beta &= Kr/R_1R_2R_3 \\ \text{phase of } A\beta &= \phi-\theta_1-\theta_2-\theta_3 \end{aligned}\right\} \qquad (9)$$

If therefore X is on the root-locus so that the magnitude of $A\beta$ is unity and the phase is 180° (or any odd multiple of 180°), we have

$$\theta_1+\theta_2+\theta_3-\phi = \text{odd multiple of } 180° \qquad (10)$$

$$K = R_1R_2R_3/r \qquad (11)$$

Now the essential point is that in order to plot the root-locus, we have to find positions of X such that Equ. (10) is true, and we can do this by trial and error, and by measurement of the distances and angles in Fig. 1 if we wish, or by calculation if we prefer it or need accuracy greater than is obtainable by plotting and measurement.

The easiest points of the root-locus to find are those for which X is on the real or α-axis. At such points the angles θ_2 and θ_3 will necessarily add up to zero (or 360°) so that Equ. (10) effectively reduces to the fact that $(\phi-\theta_1)$ is an odd multiple of 180°. Now if X is on the real axis to the left of P_1, both ϕ and θ_1 will be 180°, and if X is to the right of Z and on the real axis, both ϕ and θ_1 will be zero, so X can only be on the real axis between P_1 and Z, when θ_1 is zero but ϕ is 180°. In fact, the portion P_1Z is a complete branch of the root-locus; if K is small, there will be a root near P_1 while if K is large, there will be a root near Z.

Since when $K = 0$ there is a root at P_2, there must be a branch of the root locus starting at P_2, and our next task is to find out what happens to this branch. If X is near P_2, θ_2 will be very sensitive to small variations in the position of X, but the angles θ_1, θ_3 and ϕ_3 will differ little from their values when X is actually at P_2, namely $\theta_1 = \tan^{-1}5 = 78°41'$; $\theta_3 = 90°$ and $\phi_1 = 180° - \tan^{-1}5 = 101°19'$. Hence from Equ. (10), the branch beginning at P_2 leaves P_2 so that $\theta_2 = 112°38'$. Further, if X is a long way off in a direction making an angle θ with the real axis $O\alpha$, then θ_1, θ_2, θ_3 and ϕ will all be equal to θ, and therefore to satisfy Equ. (10) we must have $\theta = 90°$ or 270°. The branch of the root-locus which starts at P_2 therefore goes to infinity along an asymptote which is parallel to the imaginary or ω-axis. More accurately, if X is a long way off in the direction of the imaginary axis, $\tan\theta_1 = \omega/(\alpha+4)$, and for large ω this means that θ_1 is $90° - \tan^{-1}\{(\alpha+4)/\omega\}$ or $\{90 - 180(\alpha+4)/\pi\omega\}$ degrees. Similarly, $\tan\theta_2 = (\omega-5)/(\alpha+3)$, and for large ω this means that θ_2 is $90 - \{180(\alpha+3)/\pi(\omega-5)\}$ degrees, and likewise θ_3 is $90 - \{180(\alpha+3)/\pi(\omega+5)\}$ degrees, and ϕ is $90 - \{180(\alpha+2)/\pi\omega\}$ degrees. If we neglect $1/\omega$ altogether, we obtain the result we have already had, that X must go to infinity parallel to the imaginary axis; if however, we retain $1/\omega$ but neglect $1/\omega^2$, which means that in the denominators of the above expressions for θ_1, θ_2, θ_3 and ϕ we regard ω, $\omega+5$ and $\omega-5$ as all equal to ω, we find that to satisfy Equ. (10)

$$\frac{180}{\pi\omega}[(\alpha+4)+(\alpha+3)+(\alpha+3)-(\alpha+2)] = 0 \qquad (12)$$

so that α is -4. Hence the branch of the root-locus beginning at P_2 glides on to an asymptote parallel to the imaginary or ω-axis and a distance 4 to the left of it. (The fact that this asymptote goes through P_1 is an accident.) The complete root-locus is therefore as in Fig. 2. In this case there is no point in taking much trouble to determine any additional points, but if it was for any reason necessary to do so, α can be taken to have successive values say $-3{\cdot}3$ and $-3{\cdot}6$ and the corresponding values of ω for X to be on the root-locus can then be found by trial and error. If $\alpha = -3{\cdot}3$, we know that ω must be somewhat greater than $5 + [\alpha-(-3)]\tan 112°38' = 5.72$, since the branch of the root-locus

* Equs (7) and (8) are valid for all positions of X; it is often easiest to draw figures of this kind as if α, ω are both positive and assume that in other cases "the signs take care of themselves".

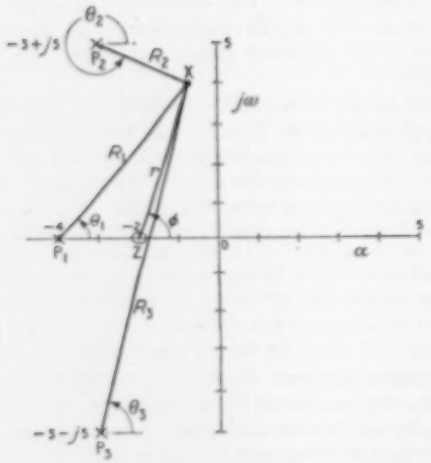

Fig. 1. *Poles and zeros associated with Equ. (6), together with a typical point X or $\alpha+j\omega$ (α negative)*

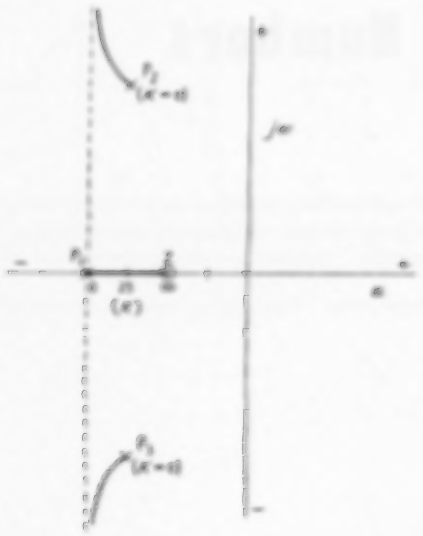

Fig. 2. *Rough sketch of root-locus associated with Equ. (5)*

we are considering starts from $-3+3j$ at the angle 112° 38′ and ends at $-4+j\infty$ and is therefore concave upwards. We therefore start with a trial value $\omega = 5{\cdot}8$, in which case

$$\left.\begin{aligned}&\theta_1{=}83°7'\ \ \theta_2{=}110°33'\ \ \theta_3{=}91°35'\ \ \phi{=}102°38'\\&\theta_1+\theta_2+\theta_3-\phi = 182°37'\end{aligned}\right\} \quad (13)$$

and the correct value of ω is in fact very close to 5·95, for when $\omega = 5{\cdot}95$, $\theta_1+\theta_2+\theta_3-\phi = 180°3'$. It is quite sufficient to obtain ω within say $\pm 0{\cdot}1$ by meaurement, because at this stage we only need to know the salient features of the root-locus. If for any reason we particularly need to know a root accurately for a particular value of K, this can be obtained by methods given in earlier 'Tools' (February and March 1957) when the starting approximation obtained by the root-locus method will be very useful. The complete root-locus for the case of Equ. (5) is thus as in Fig. 2. We already know that the branches begin at the poles when $K = 0$ and end at the zero Z or at infinity when K is large; if we require to know more accurately the value of K at any other point of the root-locus, we can obtain it from Equ. (11), and graphical accuracy will be sufficient. It is much easier to choose a point of the root-locus and determine K from Equ. (11) than to choose K and determine the corresponding root. If therefore what we really require is the approximate value of the root associated with a particular value K_0 of K, on a particular branch, we can find a point X_1 on that branch where K is slightly less than K_0, and another where K is slightly greater than K_0 and then interpolate graphically between them.

In this particular case, the equation $1+A\beta = 0$ was only cubic in p, and therefore an alternative procedure was possible, namely to choose a real negative value of p between -2 and -4 (which we already know is on the locus), determine the corresponding value of K (by calculation) from Equ. (11), and then we have only a quadratic equation to solve to find the remaining two roots associated with the same K. Thus if we take $p = -3$, $R_1 = 1$, $R_2 = R_3 = 5$ and $r = 1$, so that $K = 25$. Hence

$$1 + A\beta = 1 + 25\,\frac{(p+2)}{(p+4)\,[(p+3)^2+5^2]} \quad (14)$$

so that the equation we actually have to solve for the roots is

$$(p+4)\,(p^2+6p+34)+25\,(p+2) = 0 \quad (15)$$

knowing that one of the roots is $p = -3$. In fact

$$(p+4)\,(p^2+6p+34)+25\,(p+2) = p^3+10p^2+83p+186 = (p+3)\,(p^2+7p+62) \quad (16)$$

so that the remaining two roots for $K = 25$ are $-3{\cdot}5 \pm j\,7{\cdot}0534$. But the root-locus method is available whatever the degrees of $N(p)$ and $D(p)$ may be, whereas the method just described would not work if the equation $1+A\beta = 0$ was of high degree, and especially if most of the roots were complex.

In this article we have applied the root-locus method to a particular example, Equ. (5), which has some but not quite all of the features which commonly occur. In our next article we shall consider another case in which the missing features do occur, and the way in which the principles used in determining the root-locus of Fig. 2 can be expressed as simple rules which are generally applicable.

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1961–62 I.E.E. COUNCIL ELECTION

The following were nominated for the vacancies which will occur in the Institution of Electrical Engineers' Council on 30th September next and, as no other nominations were received, the Council's nominees were duly elected:

President, G. S. C. Lucas, O.B.E.; Vice-Presidents, A. H. Mumford, O.B.E., B.Sc.(Eng.) and R. L. Smith-Rose, C.B.E., D.Sc., Ph.D.; Hon. Treasurer, C. E. Strong, O.B.E., B.A., M.Sc. Ordinary Members of the Council: Members, Sir Robert Cockburn, K.B.E., C.B., M.Sc., Professor C. W. Oatley, O.B.E., M.A., M.Sc., J. S. McPetrie, C.B., Ph.D., D.Sc. and H. West, M.Sc. Associate Members, J. Brown, M.A., Ph.D., D.Sc.(Eng.) and P. L. Taylor, M.A.; Companion, The Rt. Hon. The Viscount Falmouth.

The following have been elected to fill the vacancies which will occur on the Committee of the Electronics and Communications Section: Chairman, R. J. Halsey, C.M.G., B.Sc.(Eng.); Vice-Chairman, J. A. Ratcliffe, C.B.E., M.A., F.R.S. Ordinary Members: V. J. Cooper, B.Sc.(Eng.), H. Davies, M.Eng., G. W. A. Dummer, M.B.E., R. Feinberg, Dr.-Ing., M.Sc., C. A. Marshall, B.Sc. and R. C. G. Williams, Ph.D., B.Sc.(Eng.).

B.E.A.M.A. ELECTRONICS BOARD

The Council of The British Electrical & Allied Manufacturers' Association has established an Electronics Board to represent that part of the B.E.A.M.A. membership directly concerned with electronic engineering. The initial membership of the Board is as follows: Chairman, L. Bagrit (Elliott Bros.); Vice-Chairman, O. W. Humphreys (G.E.C.); E. B. Banks (English Electric); A. W. Steel (A.E.I.); S. de Ferranti (Ferranti Ltd.); C. Metcalfe (E.M.I.) and W. Gregson (Ferranti Ltd.).

Slide Rule for Complex Numbers

By D. J. Whythe, B.Sc. (Eng.), A.M.I.E.E.*

SUMMARY. *A conventional slide rule can be used to multiply or divide real numbers, by adding or subtracting the logarithms of the numbers plotted on a straight line. This principle can be extended to permit the multiplication and division of complex numbers, by plotting the logarithms of the numbers on the surface of a cylinder. Numbers are then multiplied and divided by sliding and rotating the cylinder.*

The rule operates directly upon numbers plotted in either rectangular or polar form. The result is indicated in both rectangular and polar form.

A complex number, denoted by $x + iy$, is defined as a number-pair in which both x and y are real numbers and $i = \sqrt{(-1)}$. The quantity x is termed the 'real' part and iy the 'imaginary' part. The number may also be written in polar form as $Re^{i\theta}$ where $R = \sqrt{(x^2 + y^2)}$ and $\tan \theta = y/x$. R is called the 'modulus' and θ the 'argument' of the number.

The first published reference to complex numbers was probably made by Girard[1] in 1629 in connection with the roots of quadratic equations but, although Wallis pointed out in 1673 that to plot such numbers would require departure from a single dimension, it was not until 1797 that both Wessel and Gauss suggested two-dimensional representations of complex numbers on a plane. In addition to their application to physical problems, complex numbers have a 'completeness' which is lacking in the algebra of real numbers. Functions which appear distinct when only real variables are considered become related when complex variables are introduced (for example, $\cos i\theta = \cosh \theta$). Again, certain theorems are not true if numbers are restricted to be real; the statement that an equation of nth degree has n roots, for example, is true in general only if complex roots are included.

Unfortunately, formulae which are simple when expressed in complex quantities become involved when, for numerical computation, they have to be transformed to permit the arithmetical processes applicable to real numbers. For addition and subtraction it is necessary to express complex quantities in rectangular $(x + iy)$ form, whereas the polar $(Re^{i\theta})$ form is much more convenient for multiplication and division. Computations often involve several transformations from one form to the other. This article describes a slide rule which uses the complex-plane representation of complex numbers, with logarithmically-divided co-ordinates. It can be used to multiply or divide complex numbers with the same facility (and accuracy) with which a conventional slide rule can be used to multiply or divide real numbers. The result is indicated in both rectangular and polar form.

* British Broadcasting Corporation.

Principle of Slide Rule

In the same way that a conventional slide rule adds or subtracts the logarithms of real numbers plotted along a straight line, the complex slide rule operates by adding and subtracting the logarithms of complex numbers plotted on a plane.

Consider the transformation

$$\log_e (x + iy) = u + iv = \log_e R + i\theta \quad \ldots \quad (1)$$

which relates a number, referred to rectangular axes drawn on the x, y plane, with the corresponding number referred to rectangular axes drawn on the u, v plane. A straight line of constant real part equal to x_1 on the x, y plane has a locus in the u, v plane given by

$$e^u \cos v = x_1 \quad \ldots \quad (2)$$

and

$$\frac{du}{dv} = \tan v \text{ (independent of } x_1) \quad \ldots \quad (3)$$

A straight line of constant imaginary part equal to iy_1 on the x, y plane has a locus on the u, v plane given by:

$$e^u \sin v = y_1 \quad \ldots \quad (4)$$

and

$$\frac{du}{dv} = \tan\left(v \pm \frac{\pi}{2}\right) \text{ (independent of } y_1) \quad \ldots \quad (5)$$

Thus, referred to the rectangular axes of u and iv drawn on the u, v plane, all loci of points of constant real or imaginary part in the x, y plane have the same shape and can be drawn to a common template satisfying

$$e^u = \operatorname{cosec} v \quad \ldots \quad (6)$$

Loci of points of constant imaginary part are displaced relative to those of constant real part by $\pi/2$ parallel to the v axis. Since the transformation is orthogonal, loci intersect orthogonally if a ratio of 1 neper occupies the same length on the u axis as 1 radian on the v axis. It follows from Equation (1) that

$$u = \log_e R \quad \ldots \quad (7)$$

and

$$v = \theta \quad \ldots \quad (8)$$

so that the u and v axes respectively may be marked logarithmically in terms of modulus, and linearly in terms of argument.

In this way, a grid of rectangular co-ordinates in the x, y plane can be transformed into the graticule shown

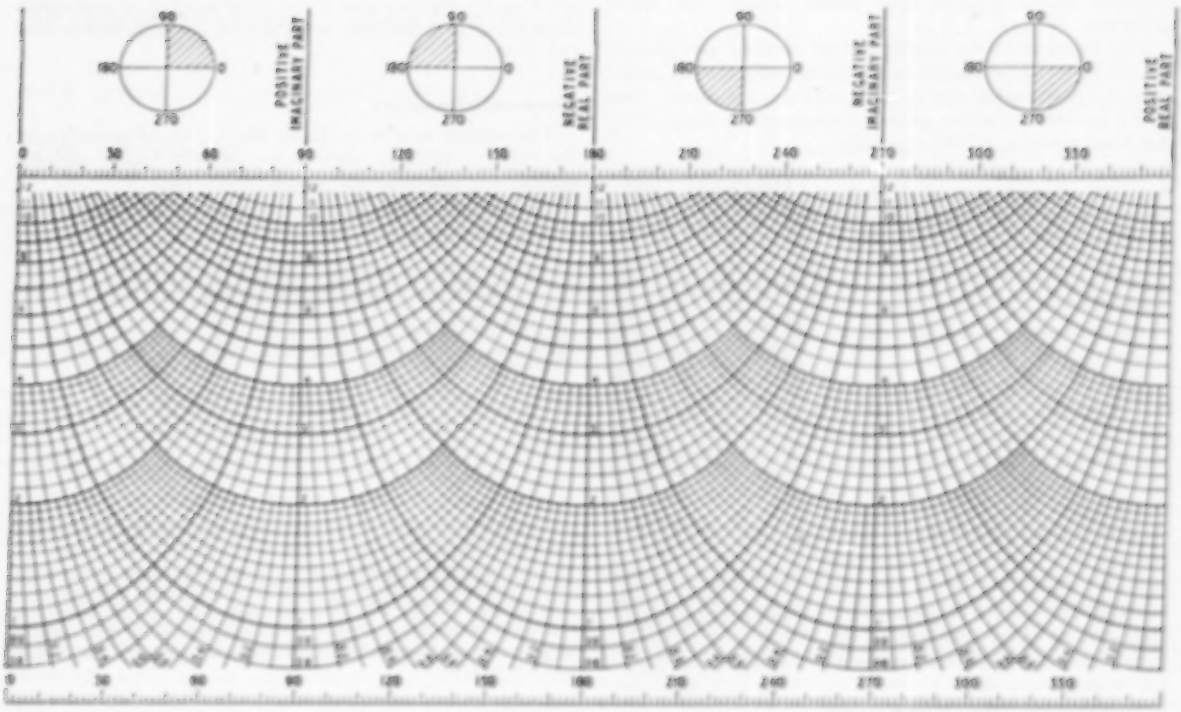

Fig. 1. *Developed form of graticule*

in Fig. 1. The graticule is divided into four equal sections which correspond to the four quadrants of the familiar Argand diagram. Numbers plotted on the graticule with reference to either the curvilinear co-ordinates of real and imaginary part, or the rectangular co-ordinates of modulus and argument, can be multiplied or divided by making the appropriate traversals parallel to the rectangular axes. The result can be read off in polar form from the rectangular axes or in rectangular form from the curvilinear co-ordinates. As an aid when using the rule, the corresponding quadrant of the Argand diagram is shown shaded above each of the four sections of the graticule.

Design

A complex slide rule using a flat graticule of the form shown in Fig. 1 has been described by DuMond[2]. The mechanism of multiplication and division is greatly simplified, however, if the graticule is engraved on the surface of a cylinder[3]. Furthermore, if the loci of numbers having argument either 0 or 2π are made to coincide, numbers are still correctly indicated if their argument is changed by any even integral multiple of π. The loci will intersect orthogonally if the length occupied by one decade on the graticule is 1·1515 times the diameter of the cylinder, but this condition is not essential and a different length/diameter ratio may be used if more convenient.

Figs. 2 and 3 show the complete rule. It embodies a movable cylinder upon which the graticule shown in Fig. 1 is engraved. Complex numbers are represented either in rectangular form by the intersection of the appropriate curvilinear loci or, in polar form, by the scales of argument marked on the cylinder together with the scales of modulus marked on two transparent cursors.

The cylinder rotates and slides on a fixed cylinder which is attached to the handle and to one of the cursors. The second cursor is attached to an inner

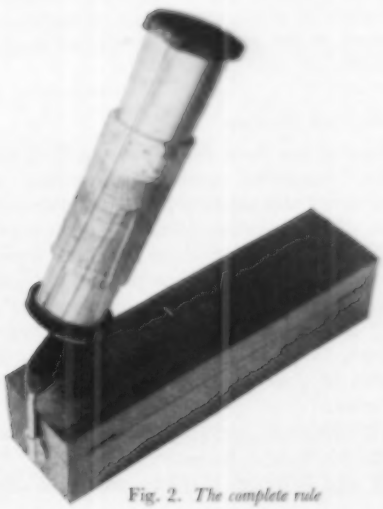

Fig. 2. *The complete rule*

cylinder which rotates and slides within the fixed cylinder.

The numbers to be multiplied or divided are plotted on the movable cylinder in either rectangular or polar form by positioning the cursors and the cylinder. The result is read in either polar or rectangular form from the final position of the cylinder relative to the appropriate cursor. As with an ordinary slide rule, the position of the decimal point must be determined by inspection.

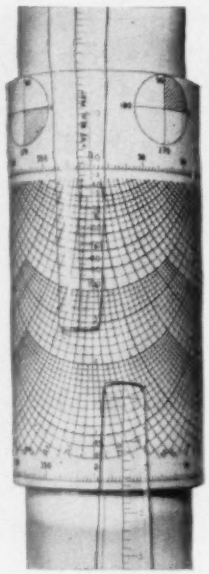

Fig. 3. *Enlarged view of graticule and cursors*

Operation

Plotting Numbers Expressed in Rectangular form

The curvilinear engravings on the movable cylinder represent the loci of complex numbers having constant real or imaginary parts. The loci are plotted at sufficiently small intervals to permit linear interpolation. Red indexes are marked on the two cursors; numbers expressed in rectangular form are plotted on the cylinder by setting one of the red indexes at the intersection of the appropriate loci.

In Fig. 3, the cursors are shown in positions which plot the numbers $(1{\cdot}4 + i0{\cdot}4)$ and $(3{\cdot}4 - i0{\cdot}4)$.

Plotting Numbers Expressed in Polar Form

Two scales of argument are engraved from 0 to 2π round the movable cylinder. Logarithmically-divided scales of modulus are engraved on the cursors; the scale on the fixed cursor is engraved from 0·8 to 11, the scale on the movable cursor from 1 to 100.

Numbers in polar form are plotted on the movable cylinder either by aligning the modulus on the fixed cursor with the lower scale of argument on the cylinder or by aligning the modulus on the movable cursor with the upper scale of argument on the cylinder.

The cursors at the positions shown in Fig. 3 plot the numbers $1{\cdot}46 \angle 16°$ and $34{\cdot}3 \angle 353°$.

Multiplication and Division

Operations of the form PQ/R (where P, Q and R are all complex quantities expressed in either polar or rectangular form) are performed by the following three movements. The movable cylinder is positioned so that the denominator R is plotted on it by the fixed cursor. The movable cursor is set so that it plots the quantity P. Finally, the movable cylinder is positioned so that the quantity Q is plotted on it by the fixed cursor. The result is read, in polar or rectangular form, from the new position of the movable cylinder relative to the movable cursor.

For straightforward multiplication or division, either of the quantities Q or R is unity.

The rule also provides a ready means of calculating $\sqrt{(a^2 + b^2)}$ where a and b are known, or of converting a complex number from rectangular to polar form and vice versa.

Acknowledgement

The author wishes to thank Mr. G. D. Monteath and Dr. G. J. Phillips who both offered many helpful suggestions towards the design of the rule.

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THORN—A.E.I. MERGER

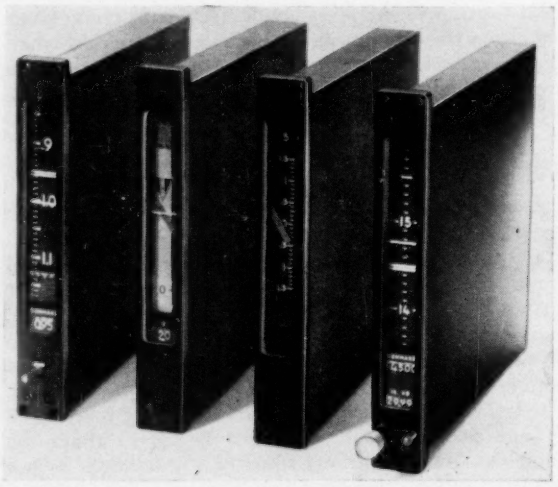
Thorn Electrical Industries and Associated Electrical Industries have announced that they are merging their respective interests in the manufacture and sale of cathode-ray tubes and radio valves. The trade marks involved include Mazda, Ediswan and Brimar.

A.E.I. and T.E.I. will both be represented on the Board and will have equal shareholdings in the joint company. Management will be vested in Thorn Electrical Industries with Mr. Jules Thorn as Chairman.

INTERNATIONAL CONFERENCE ON SEMICONDUCTORS

The Institute and Society on behalf of the International Union of Pure and Applied Physics and the British National Committee for Physics is arranging an International Conference on the Physics of Semiconductors, which will be held at the University of Exeter from 16th to 20th July 1962. Further information regarding the Conference may be obtained from the Administration Assistant, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

AIRCRAFT PANEL METER



To meet the demand for compact aircraft cockpit instruments which are easy to read, Kollsman Instrument Corporation, U.S.A. have introduced these 'bookshelf' type instruments with vertical dials. They are $1\frac{1}{2}$ *in. in width*

Waveguide Switch

FOR MICROWAVE RADIOMETERS

By M. J. B. Scanlan*

SUMMARY. *A new waveguide switch, especially useful in large waveguides, is proposed for microwave radiometers. A switch has been built in No. 6 waveguide, and some preliminary results are given.*

The radiometer, first used by R. H. Dicke[1] in 1946, is essentially a device in which a radio receiver is switched at a low frequency between two sources of radiation: any difference in power output between the two sources appears as a modulation, at the switching frequency, in the receiver output. The modulation envelope, after detection, is amplified by a narrow-band amplifier and finally applied to a phase-sensitive detector, for which the switching process provides the reference voltage. As used in radio astronomy, for which the radiometer is perhaps the fundamental instrument, the two sources of radiation are usually a matched termination or load at room temperature, on the one hand, and an aerial on the other. By the use of a radiometer, fluctuations in receiver gain, which would otherwise severely limit the sensitivity available, may be made of less importance, especially if the aerial and the load can be made to have the same noise temperature in the absence of a radio star signal[2].

Various forms of switch have been used in radiometers. At longer wavelengths, above 1 metre, devices such as capacitance switches have been used[3], while at microwave frequencies, Dicke himself used a resistive vane dipping into the waveguide: here the resistive vane itself comprises the matched termination. More recently ferrite switches have come into use[4].

There remains, however, a range of frequencies, roughly from 1,000 to 2,000 Mc/s, in which none of these techniques seems very satisfactory: this region includes the vitally important hydrogen line at 1,420 Mc/s. Coaxial-line techniques can still be used, of course, and a capacitance switch could no doubt be made, but it would require very careful design and manufacture to avoid losses and to preserve bandwidth. The waveguide devices used at higher frequencies are also unsatisfactory here: the resistive vane becomes uncomfortably large, while ferrites for these frequencies are not yet as useful as at shorter wavelengths.

A New Radiometer Switch

A new switch for use in waveguide, and especially large waveguide, was therefore sought, and a solution found which is presented here. This solution exploits the well-known properties of the short-slot (or Riblet) coupler, which consists of two lengths of waveguide, about 1 guide wavelength long, with a suitable coupling aperture in their common (narrow-face) wall. If the four ports of the two waveguides are numbered 1 to 4 clockwise so that 1 and 2 are adjacent (Fig. 1), then a signal entering port 1 appears equally at ports 3 and 4 (port 4 leading port 3 by 90° in phase), the signal at 2 being typically 30 dB down on the input. If ports 3 and 4 are closed by short circuits in the same plane, signals entering 1 all appear at 2. These relations are shown in Fig. 1.

Two such couplers are used in the switch, separated by a rotating metal vane or paddle, which alternately opens and closes the waveguides joining the two couplers. The system is shown in Fig. 2 (a) (paddle open) and 2 (b) (paddle closed): it has four waveguide ports, numbered as before. With the paddle open [Fig. 2 (a)], ports 1 and 3 are connected directly, as are ports 2 and 4: there is very weak coupling between any other pairs of ports. With the paddle closed [Fig. 2 (b)], port 1 is connected to port 2, and port 3 to port 4: coupling between port 1 and port 3, for instance, is limited to leakage through the closed switch.

It follows from this that if the paddle is rotated about

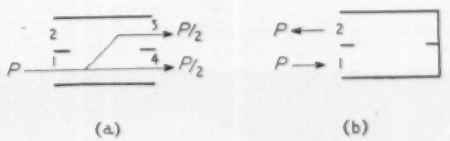


Fig. 1 (*above*). *Plan view of short-slot coupler with (a) matched loads or (b) short circuits on arms 3 and 4*

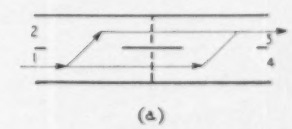
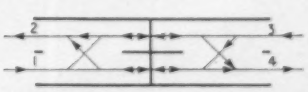


Fig. 2 (*right*). *Plan view of switch with paddle open (a), and closed (b)*



* *Marconi's Wireless Telegraph Co. Ltd., Chelmsford.*

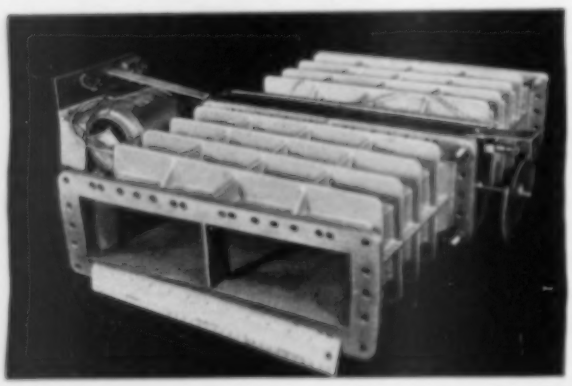
an axis through the centre of the three narrow waveguide walls, port 1 is switched between ports 2 and 3, port 2 between 1 and 4, and port 3 between ports 1 and 4. This switching is at twice the rotation frequency of the paddle, and the resultant modulation is closely sinusoidal. In use as a radiometer, the aerial could be connected to port 1, for instance, a matched termination (the reference source) to port 4, and the receiver to port 3. Port 2 can be matched, or left open—or short-circuited, since it is only very weakly coupled to any of the ports in use. Alternatively, a second receiver may be connected to port 2, and is then switched between the aerial and the reference source, just as is the first receiver on port 3, but in reverse phase. This two-receiver system can avoid one disadvantage of previous radiometers, namely that half the signal received by the aerial is lost.

Since the receiver noise signals are eventually converted to a low-frequency modulation voltage, it seems possible to subtract (because of their phase difference of 180°) the two modulation voltages from each other to give an improved signal-to-noise ratio. This possibility of using all the available aerial signal requires a four-port switch and is believed to be novel. Some other switches could doubtless be made to use all the signal available from the aerial, but the rotating resistive vane type, of course, necessarily uses only half the aerial signal. In two receiver radiometers such as that under discussion, the receivers need be only roughly matched in noise factor and gain. Furthermore, if the two receivers are swept in frequency, maintaining a constant frequency difference between them, the switch can act as a spectrograph, e.g. for the hydrogen line.

Practical Details

An embodiment of the switch in No. 6 waveguide is shown in photographs, Figs. 3 and 4. Fig. 3 shows the complete switch assembly, while in Fig. 4 one short-slot coupler has been removed to show the paddle in the half-closed position. The short-slot couplers in this switch are light-alloy castings, originally developed for high-power duplexer work, the prominent external ribbing being required to withstand internal pressurization. The couplers cover the range 1,200–1,400 Mc/s, and determine the usable bandwidth of the switch. The paddle section is of brass, and the paddle is carried on a

Fig. 3. *The complete switch assembly*



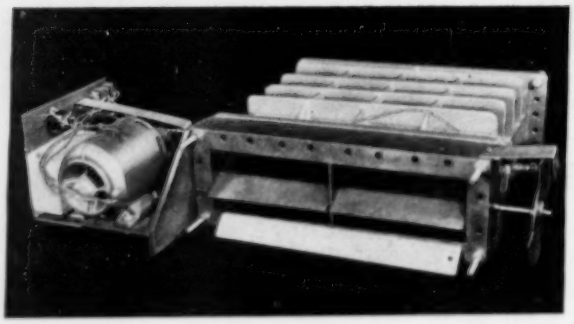


Fig. 4. *The switch with one coupler removed to show the paddle*

steel rod $\frac{1}{4}$ in. in diameter passing through the centre of the three narrow waveguide faces: no bearings are used to support the rod in this case. The rod is driven at one end by a synchronous motor, geared down to 750 r.p.m., so that the modulation frequency here is 25 c/s. The remote end of the rod carries a two-bladed chopper, which generates a reference voltage by interrupting the path between a 6-V bulb and a semiconductor photo diode. Coarse control of the reference-voltage phase is achieved by turning the chopper on its shaft, while there is a measure of fine control in operation by movement of the carriage carrying the bulb and photo diode. The choice of half mains frequency was dictated by the availability of a suitable drive motor and of narrow-band 25 c/s amplifiers, which incorporate a 50 c/s rejection filter. This choice of frequency is open to the objection that harmonics and sub-harmonics of mains frequency are better avoided: but it follows a good precedent (Dicke's radiometer used 30 c/s), and can be changed relatively easily if necessary.

Measurements on the Switch

The important feature of the switch is the bandwidth over which a good match is presented to the receivers, whatever the position of the paddle. This is illustrated by Fig. 5, which is a plot of the voltage standing wave ratio looking into arm 1 with arms 2 and 3 matched for the open and closed positions of the paddle. For a v.s.w.r. of 1·1 or less, the bandwidth is about 15% on a centre frequency of 1,300 Mc/s, dictated by the radar band for which the couplers were designed. It is unfortunate that these couplers do not cover 1,420 Mc/s, but the change of centre frequency from 1,300 to 1,400 Mc/s should present little difficulty. As for other frequency bands, short-slot coupler designs exist at 600 Mc/s, at 3,000 Mc/s, 5,000 Mc/s and 9,000 Mc/s, all having similar properties of bandwidth and directivity: the extension of the paddle sections to these frequencies should be relatively simple.

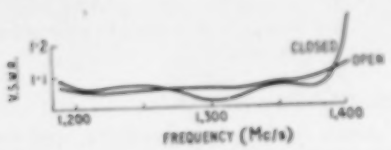


Fig. 5. *The v.s.w.r. against frequency of the switch in open and closed positions*

Insertion-loss measurements have been made, and show that the loss between ports 1 and 3 (switch open) is 0·1 dB, and between 1 and 2 (switch closed) is 0·2 dB.

One difficulty with the paddle design should be mentioned as not yet completely solved. The switching ratio (i.e., the coupling between arms 1 and 3 or 4 with the paddle closed) has the disappointingly small value so far of about 18 dB. This is a question of how closely the paddle can be made to fit its aperture, which in turn depends on how accurately the paddle shaft can be centralized in the narrow walls of the waveguide. There is room for improvement in these directions in the present switch, but to achieve better than 20 dB switching ratio will probably always require great care.

No use has yet been made of the switch, partly because of these difficulties with switching ratio and partly because a phase-sensitive detector has yet to be built. The switch is presented here as an attempt to solve some of the difficulties of radiometry in waveguide, especially where the large size of the waveguide precludes previous methods.

Conclusion

Preliminary tests on a proposed new radiometer switch show that it has at least 15% bandwidth, and that all the aerial signal can be utilized. The switching ratio is probably adequate. No tests with the switch connected to an aerial and working as a radiometer have yet been carried out.

Acknowledgements

The author wishes to thank the Engineer-in-Chief, Marconi's Wireless Telegraph Co. Ltd., for permission to publish this article, and Mr. F. E. Chittenden, who designed, and largely built, the paddle section of the switch.

REFERENCES

1 R. H. Dicke, "The Measurement of Thermal Radiation at Microwave Frequencies", *Rev. Sci. Instr.*, Vol. 17, July 1946.
2 R. Hanbury Brown and A. C. B. Lovell, "The Exploration of Space by Radio", Chapman & Hall, London, 1957, p. 37.
3 K. E. Machin, M. Ryle and D. D. Vonberg, "The Design of an Equipment for Measuring Small Radio Frequency Noise Powers", *Proc. I.E.E.*, Part III, Vol. 99, pp. 127–134, May 1952.
4 F. D. Drake and H. I. Ewin, "A Broad-Band Microwave Source Comparison Radiometer for Advanced Research in Radio Astronomy", *Proc. I.R.E.*, Vol. 46, p. 53, 1958.

MANUFACTURERS' LITERATURE

Tektronix Television Oscilloscopes. This 16-page booklet gives detailed information on three oscilloscopes which have been designed especially for maintenance and adjustment of television transmitter and studio equipment. These instruments include an oscilloscope, a waveform monitor, and the colour-television Vectorscope.
Tektronix Inc., P.O. Box 500, Beaverton, Oregon, U.S.A.

Hewlett-Packard Short Form Catalog. A 16-page catalogue giving brief details of the range of instruments produced by Hewlett-Packard and a list of their European representatives. Available in the U.K. from:
Livingston Laboratories Ltd., 31 Camden Road, London, N.W.1.

Belling & Lee Catalogue Supplement: May 1961. This 20-page supplement to the 1959–60 general catalogue has been produced to bring holders of the catalogue up to date with information concerning Belling & Lee components. Also listed are some items which are classed as "obsolescent".
Belling & Lee Ltd., Great Cambridge Road, Enfield, Middx.

Gimbal Type Flexible Joints and Bellows Assemblies. This eight-page leaflet gives details of the 'Avica-Flex' range of metal flexible joints or flexible connections in rigid tube pipelines or ducting systems.
Avica Equipment Ltd., Mark Road, Hemel Hempstead, Herts.

Silicones in Electronics. This 24-page booklet describes the main types of silicone products of use to the electronics industry. It is divided into three main sections: the first deals with encapsulating, impregnating, filling and coating materials; the second with fabricated insulating and structural components, and the third with production aids.
Midland Silicones Ltd., 68 Knightsbridge, London, S.W.1.

Pye Argon Chromatograph. In this 24-page booklet the latest Pye Argon chromatograph is described. Also included is a description of some of the techniques which have recently been developed in the field of chromatographic analysis.
W. G. Pye & Co. Ltd., Granta Works, York Street, Cambridge.

A Primer for Computing Resolvers. This 16-page booklet is designed as a simple introduction to computing resolvers, the difference between computing and data-transmission resolvers, resolver symbolism, and the methods of testing resolvers.
Theta Instrument Corporation, 520 Victor Street, Saddle Brook, N.J., U.S.A.

Indicating Pneumatic Temperature Controller. Described in this 4-page leaflet is an instrument for controlling the working temperatures of vats or other similar containers and the temperatures of refrigerating, air-conditioning, or other similar plants and processes between 340 °C and −20 °C. The instrument can be used in almost any process where the flow of heating or cooling medium can be regulated by a diaphragm valve.
Cambridge Instrument Co. Ltd., Friars Park, North Finchley, London, N.12.

The Design and Production of Printed Circuits. This 8-page booklet has been produced to act as a guide to those engaged in the manufacture of equipment using printed circuits. Various types of printed circuits are described, tolerances are set out in chart form in conjunction with a relative cost key, and electroplating specifications are given.
Technograph Electronic Products Ltd., Eros House, 29–31 Lower Regent Street, London, S.W.1.

Ediswan Industrial Valves and Cathode Ray Tubes. This 60-page booklet is the second edition of a previous publication of the same style and title. Brief details (electrical and physical) have been included on more than 30 valves and c.r.ts which have recently been added to the Ediswan industrial valve range. A list is also included of replacement and obsolete types showing the nearest equivalent current type.
Associated Electrical Industries Ltd., Radio & Electronic Components Division, 155 Charing Cross Road, London, W.C.2.

Eversheds F.H.P. Motors. A six-page leaflet which gives brief details of over 50 Eversheds units. This includes a.c. and d.c. servo motors and motor generators, special-purpose motors, tachogenerators, gearheads and torque transmission units, amplifiers and power units. The leaflet reference number is EDA 343.
Eversheds & Vignoles Ltd., Acton Lane, London, W.4.

G.E.C. Semiconductors Device Guide. This nine-page leaflet gives brief information on the electrical characteristics and physical dimensions of the range of G.E.C. semiconductors.
The General Electric Co. Ltd., Magnet House, Kingsway, London, W.C.2.

Aircraft Equipment. This 64-page illustrated booklet on A.E.I. aircraft equipment covers a.c. and d.c. power systems, ignition equipment, control, starting gear and power generation and utilization equipment.
Associated Electrical Industries Ltd., Publicity Services Dept., Rugby.

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Bridge for Displaying Complete D.C. Characteristic of Tunnel Diodes

SIR,—A simple method is presented of displaying the complete d.c. characteristic of a tunnel diode, including the negative-resistance region while simultaneously suppressing all oscillations. Diodes are normally stabilized with a shunt resistance. With high-current low-capacitance types stabilization can only be achieved if the series inductance of the diode is very small and the shunt resistance should therefore be placed extremely close to the diode terminals. Measurement of the actual diode current directly then becomes impossible and the alternative method is to measure the overall current through the combination of diode and shunt resistor subsequently subtracting the latter component. The bridge illustrated in Fig. 1 performs this subtraction automatically.

In Fig. 1 X, X_1, Y and Y_1 represent deflection plates of an oscilloscope and g_3 is the tunnel diode. The shunt resistor g_2 is chosen to be less than the minimum negative slope resistance to be expected in the diode characteristic and the bridge is balanced with the diode g_3 disconnected. (This is indicated by a horizontal trace on the oscilloscope.) The diode is then connected. The horizontal deflection on the oscilloscope gives the voltage V_{CB} across the diode. The voltage V_y across the vertical plates of the oscilloscope can be calculated in terms of the conductances G_1 and g_1, taking $G_1/G_2 = g_1/g_2$ at balance, giving

$$V_y = g_2 V_{CB}/(g_1 + g_2)$$
$$= I/(g_1 + g_2)$$

where I is the diode current. Hence for given values of g_1 and g_2, V_y is directly proportional to the current through the diode.

A typical trace is illustrated in Fig. 2. It can be calibrated by substituting a known resistor for the tunnel diode.

Another method is to make $g_1 \gg g_2$ when the voltage V_y due to the diode current will be given by

$$V_y \simeq I/g_1$$

and a calibrated oscilloscope will give values of diode current directly in terms of g_1. Sensitivity, of course, is sacrificed.

R. C. BARON.
G. H. B. THOMPSON.
D. R. HILL.

Standard Telecommunication Laboratories Ltd.,
Harlow,
Essex.
30th June 1961.

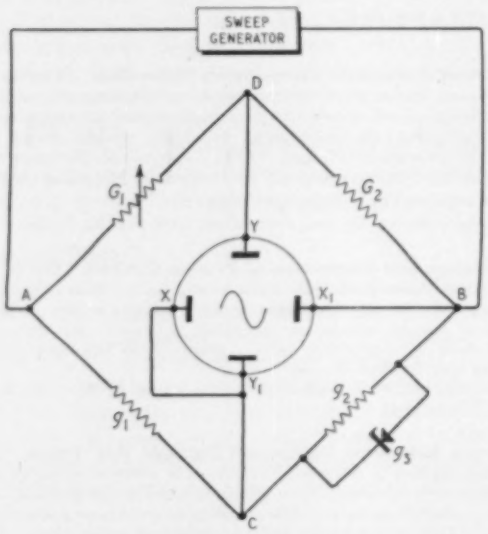


Fig. 1

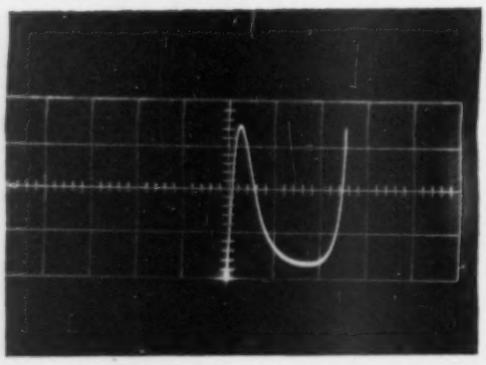


Fig. 2

The Velocity of Light

SIR,—While the recent announcement about the eclipses of Jupiter, referred to by "Quantum" in the June issue of *Electronic Technology*, is very interesting, there is nothing startling in the idea that the velocity of light may not be invariant. In the early part of the century, the only evidence in favour of invariance was the null result of the Michelson-Morley experiment, and it was shown by Ritz that this could equally well be explained if the aether were discarded and if the Galilean transformations were used, it being assumed that the velocity of light is constant with respect to the source. But for Ritz's untimely death, the development of science in this century may well have been very different.

Recent studies of my own have shown that using the above assumptions the same result is predicted for any kinematic effect as is predicted by Einstein's theory. There is also a possibility that certain effects might become tractable that have not proved so on the assumption of the invariance of the velocity of light. For example, the light curve of the eclipsing binary star Algol behaves very peculiarly, several different values of the radial component of velocity being simultaneously observable at certain times. It might be worth while examining the records to see if this effect can be reduced to order on the assumption that light beams arrive simultaneously at the earth which departed at different times, with different velocities, from Algol. This approach may also help to resolve some difficulties in interpreting light curves of Cepheid variables.

There is apparently a difficulty with the mass-energy equivalence, which is predicted by Einstein's theory and, it is claimed, verified experimentally. The key experiment was carried out by Champion in 1932. Fast β-particles from a radioactive source were first deflected by a magnetic field, in order to measure their velocities, and then allowed to collide with stationary electrons in a target in a cloud chamber. After collision, the electrons separated along paths inclined to each other at an angle dependent on the velocity of the β-particle; the relation between angle and velocity was found to agree with prediction from Einstein's theory. According to Newtonian mechanics, regarding the electrons as elastic spheres, the paths after collision would be inclined at ninety degrees, regardless of the relative velocity of the particles before collision.

Champion's result seemed to decide unequivocally in favour of Einstein's theory. Of course, one cannot disregard the facts, but at least one other interpretation is possible. In the first place, the assumption that electrons can be regarded as elastic spheres is highly questionable. Secondly, no account has been taken of the

possibility that the laws of force of electric and magnetic fields on moving charged particles may be dependent on the velocity of these particles.

In the text books, the force on a charged particle moving in a magnetic and or electric field is given as

$$\mathbf{F} = q(\mathbf{E} + \mathbf{v} \times \mathbf{B}) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

However, if the Galilean transformations are used, results consistent with Champion's observations are obtained if the force on the particle is taken to be

$$\mathbf{F} = q\left\{E\left[1 - \left(\frac{\mathbf{v}.\mathbf{e}}{c}\right)^2\right]^{3/2} + \mathbf{v} \times \mathbf{B}\left[1 - \left(\frac{|\mathbf{v} \times \mathbf{b}|}{c}\right)^2\right]^{1/2}\right\} \quad (2)$$

where **e** and **b** are unit vectors parallel to **E** and **B**. Equ. (2) reduces to Equ. (1) in the limiting case of small velocity. The apparent conflict with the basic laws of electrodynamics is resolved by noting that these laws have only been subjected to experimental test in the limiting case of small velocities. The assumption in orthodox relativity theory that Maxwell's equations hold regardless of any relative motion of the material bodies concerned is therefore not based on any experimental evidence, although it has been regarded as a fact and forms one of the cornerstones of Einstein's theory.

R. A. Waldron.

Chelmsford.
16th June 1961.

Thermionic Conversion

Sir,—In "Quantum's" article on Thermionic Conversion in your issue of July 1961, it is stated that for thoriated tungsten the '*A*' value is 60, and by approximate arithmetic the emission at 2000 °K is calculated to be 240 amp/cm².

In case your readers jump to the conclusion that for years valve makers have been working with a safety factor on emission of more than 100 to 1, I would like to point out that the generally accepted figures* for thoriated tungsten are

$$A = 3 \text{ to } 5$$
$$\phi = 2{\cdot}6 \text{ to } 2{\cdot}7 \text{ eV}$$

which correspond to a saturated emission of about 3 to 4 amp/cm² at 2000 °K, which is the average operating temperature for carburized filaments in valves.

W. H. Aldous.

The General Electric Co. Ltd.,
Hirst Research Centre,
Wembley, Middlesex.
12th July 1961.

* D. A. Wright. *Proc. Instn elect. Engrs*, 1953, Pt. III, Vol. 100, p. 125.

[Bleaney and Bleaney (Table 4.1, page 86) give $A = 60, \phi = 2.6$ eV for thoriated tungsten. They also say (p. 330) that the current density for this material is of the order 1 amp/cm² at 1900 °K which is, in fact, consistent with Mr. Aldous's figures. The arithmetic was a poor thing but my own. Quantum]

Transistor Mutual Conductance

Sir,—In your June issue, H. L. Armstrong seeks to support your advocacy* of mutual conductance as a principal parameter of transistors, presumably in place of current amplification factor.

He evidently felt it necessary to counter the objection that "transistors are current operated" by arguing that physically it is the input voltage rather than current which controls collector current. But although this may be true on the academic plane, it is hardly relevant to practical transistor circuit design, which is the usual context of the statement quoted. What is meant by it, I am sure, is that

(a) changes in transistor output current are very closely proportional to input current, whereas the relationship of output current to input voltage is conspicuously non-linear, and
(b) whereas in a valve the input resistance is normally very high, in a transistor it is quite low.

Having from the first regarded transistors as analogous to valves, with the common-emitter configuration the normal one, I am in full sympathy with exhortations, such as yours and Mr. Armstrong's, toward that way of thinking. But the best analogy is pressed too far if it introduces practical difficulties; and although in certain modes of operation Mr. Armstrong's recommendations are helpful, in the much-used small-current modes both g_m and g_{in} vary widely with V_b. The following data for the OC71 transistor illustrate the point:

V_b(volts)	g_m(mA/V)	g_{in}(mA/V)	α_e
0·10	17	0·31	56
0·15	53	0·95	56
0·20	102	1·8	56

These relate to common-emitter working, V_b being the base-to-emitter bias.

If Mr. Armstrong is anxious to carry the analogy with the valve as far as possible, and to establish that transistors are voltage operated, why has he omitted to mention the transistor's voltage amplification factor?

Incidentally, a helpful and generally understood convention is to reckon potentials with respect to the common-electrode point, rather than the collector as shown by Mr. Armstrong. This convention is readily observable in his equivalent circuit by making the voltage of the generator negative.

These critical comments notwithstanding, I enthusiastically support Mr. Armstrong and you, sir, and everyone else who tries to re-establish on the right track those who fell for the point-contact transistor and the common-base configuration and retain the same viewpoint now, to the disadvantage of present-day students—and future ones too, if relics of the false start continue to be cherished by many individuals and organizations. I especially applaud Mr. Armstrong's courage in advocating his simple equivalent circuit, when most writers and teachers give the impression that it must invariably be the full treatment or nothing.

M. G. Scroggie.

Bromley,
Kent.
4th July 1961.

* Editorial, *Electronic Technology*, March 1961, p. 75.

PAMPHONIC ACQUIRE REFLECTOGRAPH

Pamphonic Reproducers Ltd., a Pye Group company, have recently announced the acquisition of all rights in the Reflectograph Tape Recorder from Multimusic Ltd. Pamphonic will be responsible for the servicing of all existing Reflectograph recorders and will continue the manufacture of this instrument. Sales will now be handled from 17 Stratton Street, London, W.1.

SHAFT DIGITIZERS AND MECHANICAL LOW-LEVEL SWITCHES

A committee of the Electronic Engineering Association have recently produced a list of available shaft digitizers and mechanical low-level signal switches. This list includes a brief description of each item and, in most cases, the basic specification. Enquiries for copies of the list should be addressed to the Technical Secretary, The Electronic Engineering Association, 11 Green Street, London, W.1.

NEW 'COLDPOINT' DRILLS

Simms Motor & Electronics Corporation and E. D. Sassoon Banking Co. have formed a new company to manufacture the 'Coldpoint' drill.

The main features of the drill is a fluted shank with a hard metal tip and having a bore running longitudinally through the centre of the drill with outlets at the extreme tip through which a coolant is passed under high pressure. This serves not only to keep the cutting edges cool, but also to wash away swarf. It is claimed that this drill can be used on materials of any hardness at very high speeds.

New Books

Foundations of Electrodynamics

By Parry Moon and Domina Eberle Spencer. Pp. 314 + vii. D. Van Nostrand Co., Ltd., 358 Kensington High St., London, W.14. Price 73s.

This is a systematic treatment of electrodynamics developed for undergraduate electrical engineering students in M.I.T. The treatment aims to provide a precise and completely logical development of electrodynamic theory, so that the student will at all times know what assumptions are being made and appreciate their limitations. In this aim the authors have succeeded admirably.

As a preliminary, following introductory chapters on definitions of fundamentals and on mathematical operations, the nature and meaning of each of the variables contained in Maxwell's equations are carefully examined, the whole being based on the fundamental concept of charge. After a study of Maxwell's equations there follow chapters on charges with no relative motion (electrostatics) and charges with uniform motion (d.c. conduction), leading to the study of accelerated charges (alternating currents and waves). Following this there are chapters on skin effect, waves, waveguides, radiation and aerials. The final two chapters deal with relativistic electrodynamics, where they cover carefully the assumptions and weaknesses of this difficult subject.

Each chapter commences with a careful statement of the problem and its relationship to the work of the previous chapter, and terminates with a survey of what the chapter has accomplished. These provide very useful links in the arguments. The references quoted at the end of each chapter are well chosen and range widely over important publications in languages other than English—unusual in an American book.

There is some unorthodoxy of nomenclature. For instance, to distinguish between a complex quantity and its magnitude an asterisk is used. This gives rise to a rash of asterisks on the pages, and seems an unnecessary refinement. A consequence is the need for a new symbol for complex conjugate, for which a circumflex is used. This results in the complex Poynting vector appearing in the distinctly unfamiliar form

$$\mathbf{S}^* = \mathbf{E}^* \times \hat{\mathbf{H}}^*$$

Another unorthodox symbol is one, unknown to the reviewer, for the vector Laplacian. The use of hertz for c/s is certainly an unusual occurrence in a book written in English.

This is an excellent book, its outstanding feature being its clarity. The mathematical treatment is precise without being laboured; the scope is not too wide to be cumbersome. Printing and diagrams are impeccable. H.R.L.L.

Time-Harmonic Electromagnetic Fields

By Roger F. Harrington. Pp. 480 + xi. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price £5 4s. 6d.

"This book was written primarily as a graduate-level text, but it should also be useful as a reference book. The organization is somewhat different from that found in engineering books. The material is arranged according to similarity of mathematical techniques instead of according to devices (antennas, waveguides, cavities, etc.). This organization reflects the main purpose of the book—to present mathematical techniques of handling electromagnetic engineering problems. In the sense that theorems are proved and formulas are derived, the book is theoretical. However, numerous practical examples illustrate the theory, and in this sense the book is practical. The experimental aspect of the subject is not considered explicitly.

The term *time-harmonic* has been used in the title to indicate that only sinusoidally time-varying fields are considered."

Thermoelectric Materials and Devices

By Irving B. Cadoff and Edward Miller. Pp. 344 + xiii. Published by Reinhold Publishing Corporation, U.S.A. and distributed in the U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 78s.

"Recognizing the need for presenting the subject in its many facets to engineers and scientists who were interested in entering the field, the Department of Metallurgical Engineering and Office of Special Services to Business and Industry at New York University sponsored a one week 'short course' in Thermoelectric Materials and Devices, ... The chapters of this volume represent the principal lectures presented in that course. . . . The final section of the book is devoted to the theory of thermoelectric processes and the thermoelectric circuit. This is followed by a general evaluation of materials, both theoretical and experimental, for use as thermoelements. The next chapters discuss materials currently used and those being considered for future use. The last section of the book concerns the principles of device design and includes descriptions of some experimental and prototype units which have been built and evaluated."

Industry and Careers: A Study of British Industries and the Opportunities they offer

By D. E. Wheatley, M.A.(Cantab.), B.Sc.(Lond.), General Editor. Pp. 872. Iliffe Books Ltd., Dorset House, Stamford Street, London, S.E.1. Price 55s. (postage 2s. 6d. extra).

"In the main body of the book each industry is thoroughly analysed: its size, location, organization, managerial structure; the materials, techniques and plant it uses, and its attitude towards research, its personnel at every level, their work and their skills. These sections in fact constitute a unique textbook on the structure of British industries.

This detailed study of each industry is followed by a review of its career prospects: how it may be entered; what educational and other qualifications are necessary for advancement, and how they can be obtained; what training schemes, if any, there are for technologists and management."

Problems in Electronics with Solutions (2nd Edn)

By F. A. Benson, D.Eng., Ph.D., A.M.I.E.E., M.I.R.E., F.I.E.S. Pp. 249 + xv. E. & F. Spon Ltd., 22 Henrietta Street, London, W.C.2. Price 42s.

"For the second edition of this textbook new problems have been added, dealing with mesh and nodal analysis, Laplace transforms, transistor equivalent circuits, transistor amplifiers and oscillators, the use of matrices in solving network problems, ionization coefficients in gases, secondary emission by positive-ion bombardment of a cathode, Townsend's theory, breakdown, probe measurements and theory, electron temperature, concentration and mobility in a plasma, collision cross-section and scattering and amplification in a gas."

Fundamentals of Electronics

By Matthew Mandl. Pp. 574 + xvi. Prentice-Hall International Inc., 28 Welbeck Street, London, W.1. Price 50s.

" . . . a single-volume book covering the basic theory and the practical aspects of basic electronics, so that a solid foundation may be laid in preparation for more advanced or specialized studies. The aim has been to present a medium-level text specifically written for those who plan a career in such branches of electronics as digital and analog computer systems, industrial control, automation, radar, microwave electronics, radio and television and other allied fields."

Electronic Maintainability. Vol. 3

Edited by F. L. Ankenbrandt, Major General, U.S.A.F., Ret. Pp. 312 + viii. Published by the Engineering Publishers and distributed in the U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 80s.

"This book, based on the Third Electronic Industries Association Conference on Maintainability of Electronic Equipment, contains an authoritative discussion of the developments which have been changing maintainability from the vaguely practised art of yesterday to the rigorous science of today and tomorrow.

Lewis's Lending Library Catalogue Supplement 1957-1959

Pp. 306. Published by H. K. Lewis & Co. Ltd. P.O. Box 66, 136 Gower Street, London, W.C.1. Price 10s. 6d. (To subscribers to the Library 5s.).

NATIONAL BUREAU OF STANDARDS

Units of Weights and Measures (United States Customary and Metric)—Definitions and Tables of Equivalents.
Pp. 20. N.B.S. Miscellaneous Publication 233. Price 40 cents.

Atomic Energy Levels in Crystals
By JOHN L. PRATHER. Pp. 84. N.B.S. Monograph 502. Price 60 cents.

"This monograph presents a theoretical study of the sharp line absorption spectra of crystals. It is an exhaustive collection of the results of group theory that apply to the theory of atomic energy levels in crystals, together with a detailed development of the general theory to permit an intelligent application of the results to experimental problems."

Amplitude-Probability Distributions for Atmospheric Radio Noise
By W. Q. CRICHLOW, A. D. SPAULDING, C. J. ROUBIQUE and R. T. DISNEY. Pp. 22. Price 20 cents.

These publications are obtainable from Superintendent of Documents, U.S. Government Printing Office, Washington 25, D.C., U.S.A. An additional one-fourth of the publication price should be added to the publication price to cover postage outside U.S.A.

BRITISH STANDARDS

British Standards Yearbook 1961
Pp. 630. Price 15s.

This Yearbook contains information about the Institution and the services it offers. It lists the British Standards current at 1st January 1961 and gives a brief description of each.

It also contains lists of publications of I.S.O., the International Organization for Standardization, I.E.C., the International Electrotechnical Commission, and C.E.E., the International Commission on Rules for the Approval of Electrical Equipment, with brief descriptions of each.

Shields for Electronic Tubes and Valves, Part 1: General Requirements and Tests.
B.S. 2258: Part 1: 1961. Pp. 8. Price 5s.

"This new British Standard applies to shields for electronic valves of the types included in B.S. 448, 'Electronic-Valve Bases, Caps and Holders'. Its scope is limited to valve shields compatible with valveholders of the types included in B.S. 448 and in a British Standard for valveholders for telecommunication and allied electronic equipment, which is in course of preparation."

Normal Equal-Loudness Contours for Pure Tones and Normal Threshold of Hearing under Free-Field Listening Conditions
Pp. 14. B.S. 3383: 1961. Price 4s. 6d.

". . . this new British Standard follows closely the draft recommendations prepared by the Acoustics Committee of the International Organization for Standardization (ISO).

Curves defining relations between frequencies characterizing sounds and their sound pressure levels for the condition of constant loudness, are involved in many aspects of subjective acoustics."

Electronic-Valve Bases, Caps and Holders
B.S. 448: Section B8D/3. Pp. 8. Price 3s. Section 7/3. Pp. 8. Price 3s.

These two sections of B.S. 448 lay down dimensions for valveholders and for the gauges needed to ensure compatibility with the appropriate valves. Section B8D/3 applies to holders, and associated gauges, for valves having B8D type bases. Section 7/3 applies to holders, and associated gauges, for valves having 5 or 7 lead-in-line bases.

These publications are obtainable from the British Standards Institution, 2 Park Street, London, W.1.

MEETING

The Institute of Navigation

26th September. "The Transit Navigation Satellite System". Papers to be presented by Dr. R. Kershner and Dr. R. Newton and others. This is a whole-day meeting to be held jointly with the Electronic Engineering Association, commencing at 10 a.m. at the Royal Geographical Society, 1 Kensington Gore, London, S.W.7.

BRITISH INSTITUTION OF RADIO ENGINEERS

Her Majesty the Queen has been pleased to approve the grant of a Charter to the British Institution of Radio Engineers. The date of the Order in Council is 2nd August 1961.

Admiral of the Fleet The Earl Mountbatten of Burma, K.G., is named in the Royal Charter as Charter President of the Institution. Graham D. Clifford is named as the first Secretary of the Chartered body; he has been Secretary of the Institution since 1937.

COLOUR TELEVISION COURSE

A refresher course of six lectures will be given on the six consecutive Monday evenings between 18th September and 23rd October 1961 in the large Lecture Theatre at the London School of Hygiene and Tropical Medicine. The course of lectures will cover a description of the N.T.S.C. system, transmission and receiver problems. The lecturers will be Mr. S. N. Watson, B.B.C. Designs Department; Mr. G. B. Townsend and Mr. P. Carnt of the G.E.C. Hirst Research Centre.

Application for enrolment forms should be made to the organizers: The Television Society, 166 Shaftesbury Avenue, London, W.C.2.

J.L.T. INDUSTRIAL RESEARCH FELLOWSHIPS

J. Langham Thompson Ltd. have announced the introduction of a scheme of industrial research fellowships to be granted on a two-year tenure for research within certain broad fields determined by the activities of the Company. These fields include electronics, cybernetics, optics, circuit techniques, logical systems, transducer techniques and mechanics of solids.

"MEASUREMENT AND CONTROL"

A new monthly journal entitled *Measurement and Control* is to be launched next January by Iliffe Production Publications Ltd. Editor-in-Chief is P. J. Farmer, A.F.R.Ae.S., who will continue as editor of the Iliffe quarterly *Data Processing*.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for July* 1961

Date 1961 July	MSF 60 kc/s 1500 U.T. Parts in 10^9	Droitwich 200 kc/s 1030 U.T. Parts in 10^9
1	− 14·9	+ 11
2	− 14·6	+ 10
3	− 15·1	+ 11
4	− 15·0	+ 11
5	− 15·0	+ 12
6	− 15·0	+ 11
7	− 15·1	+ 9
8	− 15·0	+ 7
9	− 15·0	+ 7
10	− 15·1	+ 7
11	− 15·1	+ 8
12	− 15·0	+ 9
13	− 15·1	+ 8
14	− 15·1	+ 9
15	− 14·9	+ 10
16	− 14·9	+ 10
17	− 14·8	+ 8
18	− 15·2	+ 9
19	− 15·2	+ 9
20	− 15·2	+ 8
21	− 15·0	+ 8
22	− 15·0	+ 7
23	− 14·9	+ 6
24	− 14·8	+ 6
25	− 14·7	+ 6
26	− 15·1	+ 4
27	− 14·9	+ 5
28	− 14·9	+ 4
29	NM	NM
30	NM	NM
31	NM	+ 4

* *Nominal frequency corresponds to a value of* 9 192 631 770 *c/s for the caesium F,m (4,0)—F,m (3,0) transition at zero field. NM = Not measured.*

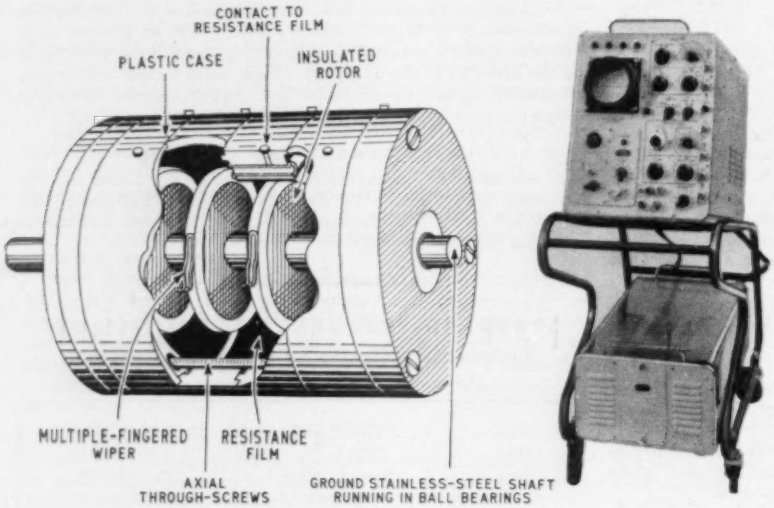
New Products

Precision Potentiometers

A range of precision potentiometers which incorporate CIC-SuperCon Film elements are now available in the U.K. through CCL Computer Controls.

Manufactured by Computer Instruments Corporation, U.S.A., these potentiometers use multiple-fingered wipers tracking on a film resistance element which is integral with the interior periphery of the case.

It is claimed that these potentiometers have virtually infinite resolution and a low noise value which is constant over the entire resistance track and which does not increase with the speed of rotation. They are capable of operation at 2,000 r.p.m. and have a life rating of a minimum of 3×10^7 operations at 100 r.p.m.



Linear models to a linear accuracy of 0·01% and sine-cosine models to a conformity of 0·025% are available. There is also a range of rectilinear potentiometers up to 16 in. in track length. The smallest of the units is the size 05 model 51 measuring ½ in. in diameter and available down to 0·2% linearity at 500 Ω. The design is such that elements can be ganged together and accurately aligned or phased.

Distributed in the U.K. through

CCL Computer Controls Ltd.,
19 *Buckingham Street, London, W.C.2.*

High-Speed Oscilloscope

Airmec have recently introduced an oscilloscope which has been designed primarily for high-speed pulse applications.

Known as type 294, it incorporates a Y amplifier with a bandwidth of d.c. to 50 Mc/s and a rise time of 7 mμsec. The sensitivity is variable from 25 mV to 10 V/cm by a 9-position attenuator which has an accuracy of calibration of not less than ±3%. Two time-base generators provide a range of sweep speeds calibrated from 2 sec. to 5 mμsec/cm with an accuracy of not less than ±3%.

Full 'M-scope' (strobe) facilities are incorporated, enabling relative time measurements to be made with an accuracy approaching 1% and permitting expansion of a selected portion of a waveform up to $\times 10^6$; two alternative methods of triggering ensure completely jitter-free strobe display under all operating conditions. Facilities for single-shot operation are provided. Initiation of the sweep may be manual or, alternatively, the sweep may be armed manually and subsequently fired by the first signal following.

The time bases may be synchronized either internally from the Y input or the 50-c/s mains supply, or from external sources; positive or negative polarity may be selected and the time base can be operated in either the repetitive or triggered mode; in the latter case there is provision for variation of the triggering threshold over a range of ±20 V.

Four auxiliary outputs are available at sockets on the front panel; these are: (a) a positive rectangular pulse of 40-V amplitude, derived from the slow sweep bright up and equal in duration to that sweep; (b) a similar pulse derived from the fast sweep; (c) a positive triangular pulse of 15-V amplitude and about 1·5 μsec duration, derived from the strobe-delay circuit and occurring at the start of the strobe; and (d) a calibrating square wave generated by an independent circuit.

Airmec Ltd.,
High Wycombe, Bucks.

Symmetrical Transistor

The Ediswan Mazda Type XS104 is a symmetrical transistor which has been introduced to meet the demand for a symmetrical device where the very tight control parameters embodied in the Ediswan Mazda Type XS101 are not necessary.

The XS104 is a symmetrical germanium p-n-p junction-type transistor, suitable for switching and modulator applications. The two p-type elements are identical and, with appropriate bias conditions, either will act as emitter with the other as collector. The average characteristics for the two conditions are identical.

At an ambient temperature of 45 °C, the maximum peak or mean collector–base voltage is −21 V and the maximum dissipation in free air is 90 mW. In the common-emitter configuration, the switching characteristics are: collector–emitter saturation voltage ($I_c = -100$ mA, $I_b = -15$ mA) $= -0{\cdot}1$ V (average)*; base–emitter forward voltage ($I_c = -100$ mA, $I_b = -15$ mA) $= -0{\cdot}5$ (average)*; and static-current amplification ($I_c = -100$ mA, $V_{ce} = -1$ V) $= 6{\cdot}5$ (min.)* and 18 (average)*.

At an ambient temperature of 25 °C (common emitter; $V_c = -5$ V, $I_c = -1$ mA) the common-emitter current amplification (β) is 10 (min.)* and 20 (average)* and the common-base cut-off frequency is 2 Mc/s (min.).

Maximum spread of β ratio (degree of asymmetry) is 3 : 1.

* Values quoted represent limits of production spreads

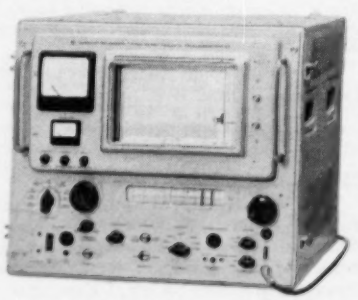
Associated Electrical Industries Ltd.,
Semiconductors Department,
Radio & Electronic Components Division,
155 Charing Cross Road, London, W.C.2.

A.F. Spectrum Analyser and Recorder

The Rohde & Schwarz audio-frequency spectograph Type FNA, recently introduced into the U.K. by Aveley Electric, has been designed for quick and accurate amplitude response and spectrum analysis of a.f. equipment.

This instrument resolves alternating voltages irrespective of waveform, making possible the analysis of pulse voltages, sound, vibration, etc. Harmonic analysis and intermodulation measurements are provided for the determination of harmonic and intermodulation distortion and modulation products.

An automatic frequency sweep allows all measured values to be plotted by the built-in recorder, the charts of which provide convenient permanent records.



The Spectograph Type FNA has a frequency range of 30 c/s to 20 kc/s; it has extreme selectivity which retains its absolute value over the entire frequency range and provides facilities for both linear and logarithmic indication.

Distributed in the U.K. by

Aveley Electric Ltd.,
Ayron Road, Aveley Industrial Estate,
South Ockendon, Essex.

Decade Resistance Box

An inexpensive resistance box which uses high-stability cracked-carbon resistors and has an accuracy of ±1% is now being produced by Levell Electronics.

This is the Type R410 and it covers the range 10 Ω to 111,100 Ω in 10-Ω steps. With each of the four decade switches set at zero the resistance ('zero resistance') is approximately 20 mΩ. At 43°C the power rating is

3 W and at 70°C 2 W; noise level is less than 0·2 μV per volt (d.c.) applied.

The resistance box measures 9¼ in. × 4 in. × 4 in. and weighs 2 lb.

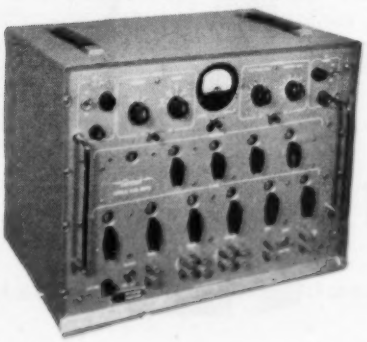
Levell Electronics Ltd.,
10–12 *St. Albans Road,*
Barnet, Herts.

Complex Ratio Bridges

Wayne Kerr have announced two Gertsch complex ratio bridges which have been designed for the precise measurement of in-phase and quadrative-voltage ratios of three- and four-terminal networks. The instruments also measure the phase-shift angle between 0 and 1° or 0 and 10°, and the tangent of the phase-shift angle.

Having an accuracy of 0·001%, six-place resolution and needing no external calibration sources, the bridges operate over the frequency ranges 30–1,000 c/s (type CBR-1B) and 50–3,000 c/s (type CBR-1B).

The basic circuit is the same in both



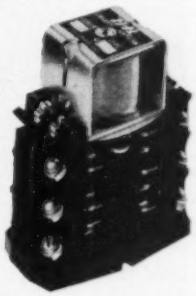
instruments; a ratio transformer functions as the ratio arms of an a.c. bridge and the off-null voltage obtained during the balancing operation is fed to a high-gain (× 500) amplifier and an indicating meter; the resulting sensitivity of the null detector is 10 μV.

Both instruments have a power-input rating of 100 W at 115–230 V, 50–60 c/s and when supplied complete in a cabinet they each measure 19½ in. wide × 16½ in. high × 13 in. deep.

Wayne Kerr Laboratories Ltd.,
44 Coombe Road,
New Malden, Surrey.

Industrial Contactor/Relay

B & R Relays have recently produced the first true contactor relay in their range. It is called the Q51 and is for a.c. operation as a heavy-duty switch or as a normal relay.



There are two versions of the relay, the Q51/2 with two switch banks and the Q51/3 with three banks; both are suitable for 440 V, 3-phase, 50 c/s operation with the contacts rated at 10 A. Operating speeds on the 2-bank version are 10–30 msec for pull-in and 10–25 msec for drop-out.

The Q51 is designed to withstand a flash test of 2,000 V and a variety of normally-open, normally-closed or changeover contacts are possible. The 2-bank version weighs 11 oz. and the 3-bank 13 oz.

B. & R. Relays Ltd.,
Temple Fields, Harlow, Essex.

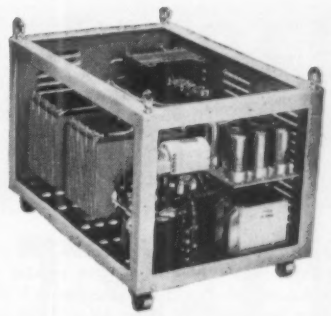
50A Stabilized D.C. Supplies

Two units which have been designed to provide stabilized d.c. at relatively high power and with high response speed to changes in supply voltage and load current have recently been announced by Claude Lyons.

Known as types PSS-32/50-A and B, they each incorporate a circuit in which the a.c. supply to the rectifiers is adjusted automatically by a motor-driven Variac, which is controlled by a transistorized servo mechanism so as to correct for any error in the pre-set output voltage.

Each model has a continuous output current rating of 50 A; type PSS 32/50-A has an output voltage range from 24 to 32 V d.c., in one continuous range; and type PSS 32/50-B provides an output range of 20 to 32 V d.c. in two overlapping ranges of 20 to 28 V and 24 to 32 V.

The required a.c. input for the standard models is 240 V −10%, 47–65 c/s single phase and the stability is ±0·5%, after



correction, against input voltage variations of ±10% and from zero to full load. The correction speed for a 20-V mains input change is less than one second.

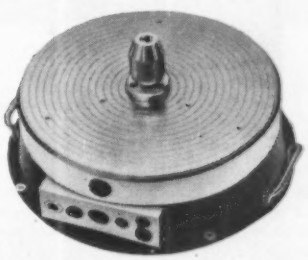
Claude Lyons Ltd.,
Valley Works,
Hoddesdon, Herts.

Polar-Recording Turntable

B & K have recently introduced a polar-recording turntable which has been designed primarily for use in conjunction with their level recorder type 2305.

The unit is intended to hold and rotate such devices as microphones, loudspeakers and underwater transducers so that, with additional appropriate arrangements, the polar responses of the devices can be plotted.

The turntable is made up of two main elements, a base and a top plate. The top plate is rotated by a synchronous motor and is provided with various means for fixing specimens, including six threaded holes and a removable chuck. Designed to carry a load of up to 220 lb, the turntable measures 14 in. diameter and is fitted with slip rings so that



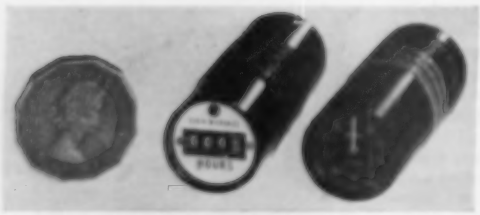
power or signals can be supplied to the specimen. In addition, it is arranged that the turntable can operate continuously or for single revolutions.

B & K Laboratories Ltd.,
4 Tilney Street, London, W.1.

Elapsed-Time Indicators

Two elapsed-time indicators which are produced by Elgin National Watch Co., U.S.A., are available in the U.K. through Aeromaritime.

The indicators have either a clock-type dial or digital read-out and will record up to 1,000 hr; units for other time ranges are also available. They are of cylindrical shape with a diameter of 0·68 in. and length of 1·42 in. (dial type) or 1·76 in. (digital type). The indicators are operated by a synchronous electric motor and a gear chain and



they are available for single-phase 400 c/s 115 V or 26 V operation; there is also a 28 V d.c. model. Power consumption is 1 W.

Over the frequency range 10 to 2,000 c/s the indicators will withstand vibrations resulting in accelerations up to 20 *g* and they will operate over the temperature range 250 °F to −65 °F. The units are filled with an inert gas and hermetically sealed.

Aeromaritime Ltd.,
45 Ladbroke Square, London, W.11.

Insulation Tester

A general-purpose insulation and continuity tester is the latest addition to the Taylor range of test instruments. This is called the Insulation Tester Model 130C and it is designed for resistance measurements from 20 Ω to 1,000 MΩ at a maximum test voltage of 500 V d.c. at infinity.

The mechanical design of the instrument takes into account the possibilities of rough handling and use by semi-skilled operators

i.e. that it is housed in a robust plastic case with a spring-mounted meter movement, with a clear 3 in. scale and simple controls. The controls comprise a three-position range and on/off switch, a zero-set control and push-button switch for the application of the test potential to the component under test.

The Model 130C operates from mains supplies of 110 V and 240 V, 40–60 c/s.

Taylor Electrical Instruments Ltd.
Montrose Avenue, Slough, Bucks.

Electronic Process Timers

Richard Allan have recently expanded their range of products by the introduction of a series of electronic timers which have been designed for short-period timing of industrial processes.

Three models cover the total timing range of 0·1 to 305 sec; type PT1L operates over the range 0·1 to 5 sec, PT1 from 1 to 55 sec, and PT1H from 1 to 55 sec plus 5 additional switched intervals of 50 sec each; i.e., 305 sec total. The basic repetitive accuracy is ±1% while the setting accuracy does not change

by more than 5% over a range of mains voltages from 200 to 250 V a.c.

The timing period is initiated by the closing of the external trigger contacts, a pulse of 40 msec duration being sufficient for this purpose. During the timing period, a pair of 5 A 250 V contacts are closed or opened. For use as a repetitive timer the trigger contacts are short-circuited and the timer will repeat the timing period with a recovery time of 40 msec.

The timer is built on the 'fail safe' principle so that interruption of the mains supply or failure of any component will open the load circuit.

Richard Allan Radio Ltd.,
P.O. Box No. 3,
Taylor Street, Batley, Yorks.

Transistor Test Meter

A general-purpose transistor-leakage test and characteristic meter is now being manufactured by Grundy & Partners. This incorporates a basic 100 μA meter movement with switched ranges of 100 μA, 250 μA, 1 mA, 10 mA and 100 mA f.s.d. A two-position switch is fitted to enable n-p-n. or p-n-p. devices to be tested and, in addition, two sets of terminals and a throw-over switch are provided to facilitate the comparison of two similar transistors.

With this meter it is possible to measure collector–emitter and collector–base leakage current respectively at a potential of 4·5 V. Also, current gain can be measured for transistors handling up to 300 mW and a reasonable indication is given for higher-powered transistors; the collector–emitter potential being fixed at 4·5 V. The base input current can be varied through a pre-set range of 10, 50, 100 and 500 μA; a push-button switch is also provided to increase the selected base current by $\frac{1}{10}$th. Diodes can also be checked with this unit.

The overall size of the instrument is 11 in. × 7¼ in. × 3¼ in. and the case is made from fibre-glass.

Grundy & Partners Ltd.,
3 The Causeway,
Teddington, Middx.

Dual-Control Recycling Timer

Industrial Timer Corporation have announced the development of a recycling timer called the Series Dual-Trol, which is suitable for the control of virtually any operation where 'on' and 'off' intervals must be frequently changed.

The Dual-Trol comprises two timing modules, one producing a series of 'on'

electrical pulses and the other a series of 'off' pulses; in both modules the frequency of the occurrence of a pulse can be adjusted over a wide range. A total of 10 timing modules is available with the maximum time ranges extending from 6 sec to 3 hr.

The timer will operate from mains supplies of 115 or 230 V at 50 or 60 c/s and it measures 7¾ in. × 5¾ in. × 5⅝ in.

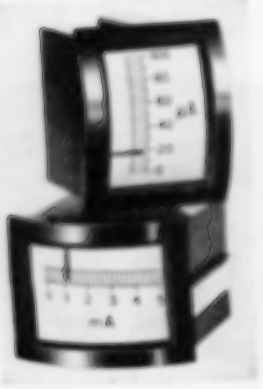
Industrial Timer Corporation,
1407 McCarter Highway,
Newark 4, N.J., U.S.A.

Miniature Edgewise Meters

A series of miniature edgewise meters, with either vertical or horizontal-plane 1 in. scales, has recently been introduced by Measuring Instruments (Pullin) Ltd. They measure 1⅝ in. × 1⅞ in. with a back depth of 2 in. and weigh about 2¼ oz.

Known as the M.I.P. Series 10, these meters are available for various current and voltage ranges from 50 μA to 500 mA and up to 250 V.

Measuring Instruments (Pullin) Ltd.,
Electric Works,
Winchester Street, London, W.3.



A BRILLIANT NEW
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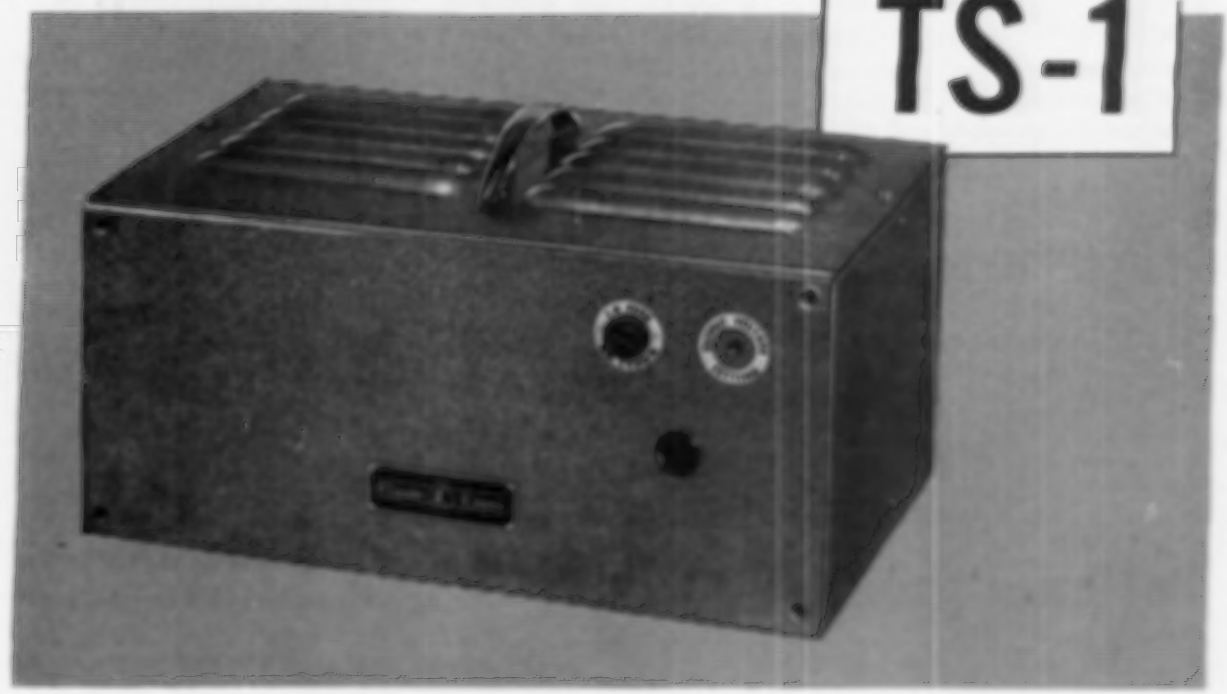

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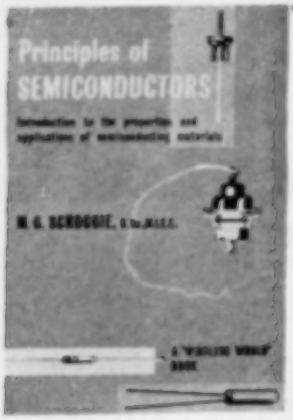

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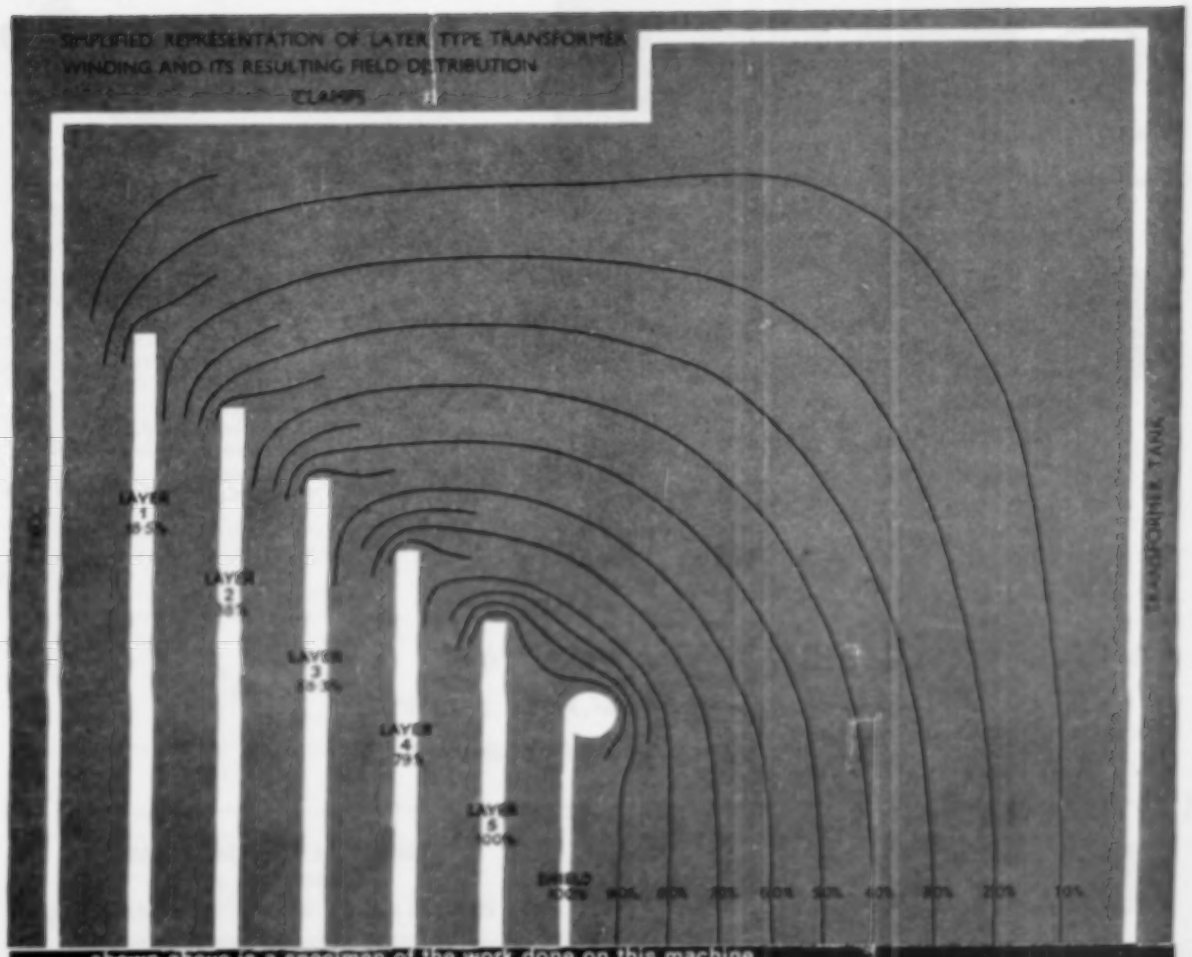

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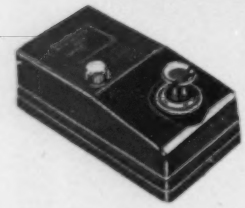

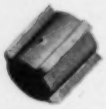

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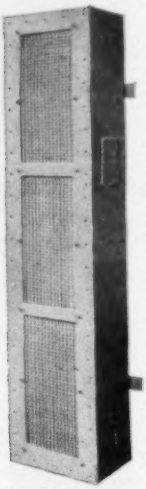

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Heater Current (amps)	I_h	0·3
Heater Voltage (volts)	V_h	7·4

TENTATIVE RATINGS AND DATA

Maximum Design Centre Ratings

		Triode	Pentode
Anode Dissipation (watts)	$p_{a(max)}$	2	1·7
Screen Dissipation (watts)	$p_{g2(max)}$	–	0·5
Anode Voltage (volts)	$V_{a(max)}$	250	250
Screen voltage (volts)	$V_{g2(max)}$	–	230
Heater to Cathode Voltage (volts rms)	$V_{h-k(max)rms}$	200	200
Cathode Current (mA)	$I_{k(max)}$	18	18

Inter-Electrode Capacitances* (pF)

Input	c_{in}	3·5	6·6
Output	c_{out}	2·1	3·1
Control Grid to Anode	c_{g-a}	1·8	0·008
Grid Triode to Grid 1 Pentode	c_{gt-g1}		0·01
Anode Triode to Anode Pentode	c_{at-ap}		0·01
Grid Triode to Anode Pentode	c_{gt-ap}		0·002
Anode Triode to Grid 1 Pentode	c_{at-g1}		0·005

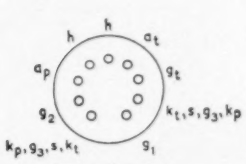
*Measured in fully shielded socket with can.

Triode Characteristics

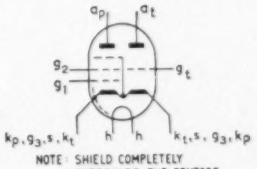
Anode Voltage (volts)	V_a	100
Anode Current (mA)	I_a	15
Mutual Conductance (mA/V)	g_m	8·5
Amplification Factor	μ	20

Base : B9A (Noval) **Mounting Position:** Unrestricted

Connections



VIEW OF FREE END



NOTE: SHIELD COMPLETELY SURROUNDS THE PENTODE.

Maximum Dimensions (mm)

Overall Length	56
Seated Height	49
Diameter	22·2

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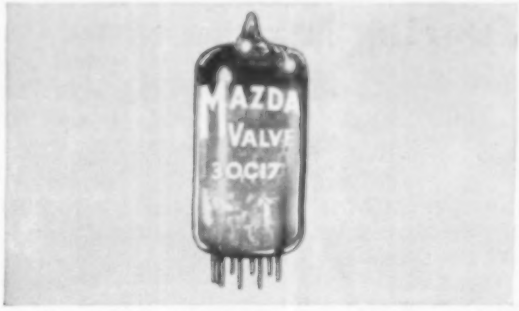

TYPICAL OPERATION AT 200 Mc/s WITH CATHODE BIAS

For operation with cathode bias where it is intended to apply AGC to the frequency changer. The oscillator voltage is applied to the pentode control grid.

Pentode

Supply Voltage (volts)	V_b	200
Anode Voltage (approx.) (Decoupling Resistance, R_a=4·7 kΩ) (volts)	V_a	170
Screen Voltage (approx.) (R_{g2}=22 kΩ) (volts)	V_{g2}	155
Cathode Bias Resistance (Ω)	R_k	100
g_1 Resistance (MΩ)	R_{g1}	4·7
g_1 Current (μA)	I_{g1}	0·6
Anode Current (approx.) (mA)	I_a	6·4
Screen Current (approx.) (mA)	I_{g2}	2·0
Conversion Conductance at 1 Mc/s (mA/V)	g_c	4·9
Grid Voltage for Conversion Conductance reduction 10 : 1 (volts)		−6·7

Triode

Anode Voltage (volts)	V_a	100
Anode Current (mA)	I_a	5

TYPICAL OPERATION AT 200 Mc/s WITH GRID CURRENT BIAS

Operation with grid current bias is suitable for the 30C17 in existing tuners not provided with AGC on the frequency changer. The oscillator voltage is applied to the pentode control grid.

Pentode

Supply Voltage (volts)	V_b	200
Anode Voltage (approx.) (Decoupling Resistance, R_a=5·6 kΩ) (volts)	V_a	148
Screen Voltage (approx.) (R_{g2}=33 kΩ) (volts)	V_{g2}	108
g_1 Resistance (MΩ)	R_{g1}	0·1
g_1 Current (μA)	I_{g1}	24
Anode Current (approx.) (mA)	I_a	9·2
Screen Current (approx.) (mA)	I_{g2}	2·8
Conversion Conductance at 1 Mc/s ($V_{het(pk)}$=2·6 V) (mA/V)	g_c	5·2

Triode

Anode Voltage (volts)	V_a	100
Anode Current (mA)	I_a	5

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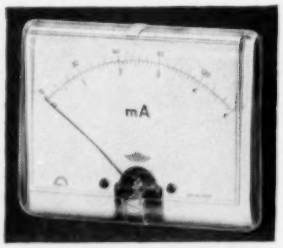

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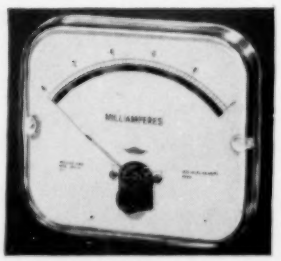

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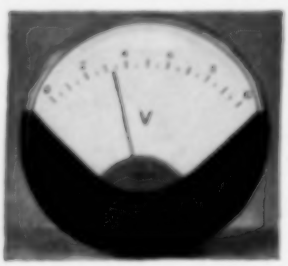

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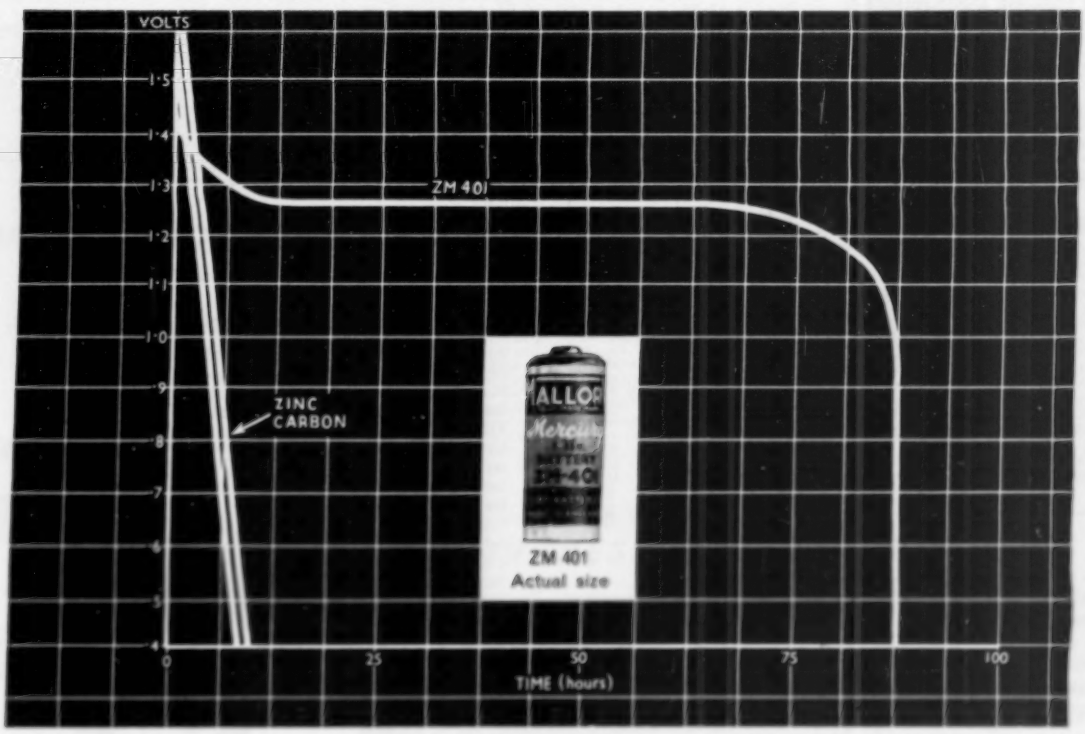

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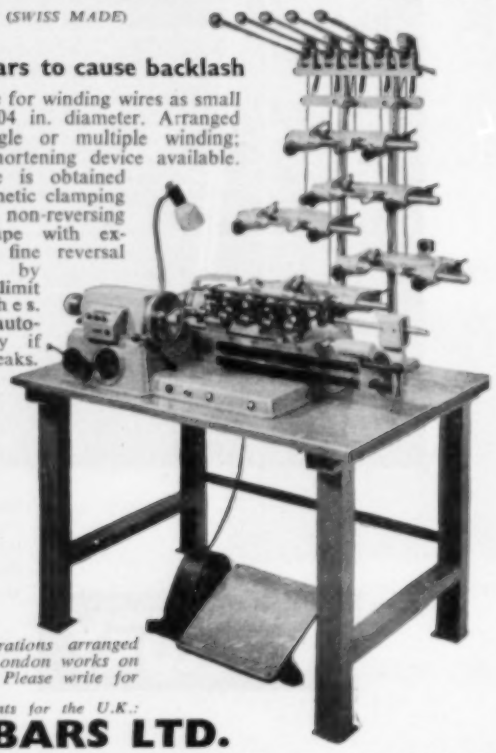

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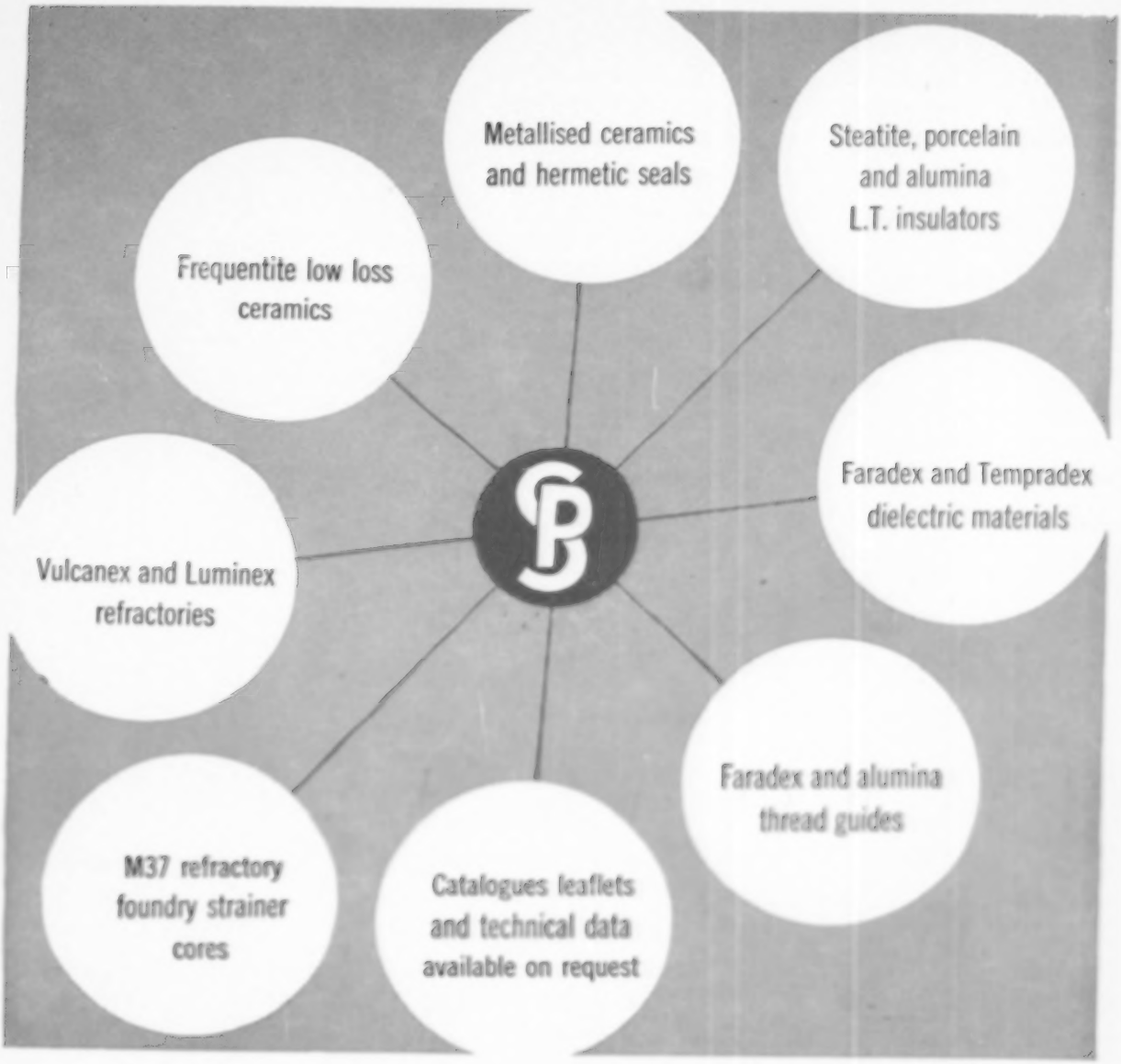
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